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Scarlatescu et al.

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(54) **FORWARD BOOST POWER CONVERTERS AND METHODS**

(71) Applicant: **Solantro Semiconductor Corp.**, Ottawa (CA)

(72) Inventors: **Gabriel Scarlatescu**, Bucharest (RO);
Raymond Kenneth Orr, Kanata (CA);
Edward Keyes, Ottawa (CA)

(73) Assignee: **Solantro Semiconductor Corp.**, Ottawa, Ontario (CA)

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H02M 3/335 (2006.01)

H02M 1/00 (2007.01)

(52) **U.S. Cl.**

CPC **H02M 3/33546** (2013.01); **H02M 3/33584** (2013.01); **H02M 3/33592** (2013.01); **H02M 2001/0058** (2013.01); **Y02B 70/1433** (2013.01); **Y02B 70/1475** (2013.01); **Y02B 70/1491** (2013.01)

(58) **Field of Classification Search**

USPC 363/16, 19, 20, 21.01, 21.02, 21.04,
363/21.06, 21.12, 21.14, 56.01, 56.09,
363/56.12, 97, 101, 124

See application file for complete search history.

(56) **References Cited**

U.S. PATENT DOCUMENTS

4,415,959 A 11/1983 Vinciarelli
6,934,167 B2 * 8/2005 Jang et al. 363/21.02
7,606,051 B1 * 10/2009 Wittenbreder, Jr. 363/39
2011/0032731 A1 * 2/2011 Coleman et al. 363/21.12

* cited by examiner

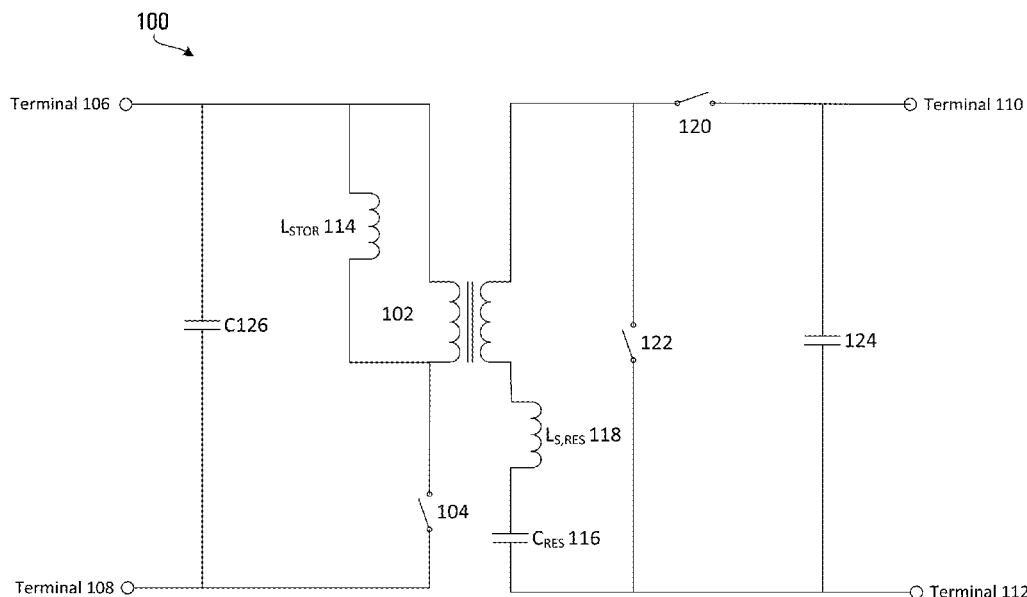
Primary Examiner — Nguyen Tran

(74) *Attorney, Agent, or Firm* — Woodard Emhardt Moriarty
McNett & Henry LLP

(57) **ABSTRACT**

Forward boost power converters, and related methods, are disclosed. In a switching mode power converter coupled between a first terminal pair and a second terminal pair, a first inductance is coupled to a first switch in a first circuit path across the first terminal pair. A capacitance is coupled to a second inductance in a second circuit path, and to the first inductance in a third circuit path. During their respective conduction periods, the first switch couples the first inductance across the first terminal pair, a second switch completes a circuit between the second terminal pair and one of: the second circuit path or the third circuit path, and a third switch completes the other of: the second circuit path and the third circuit path. Energy transfer involves both substantially linearly varying currents and substantially half sinusoidal current pulses.

23 Claims, 23 Drawing Sheets



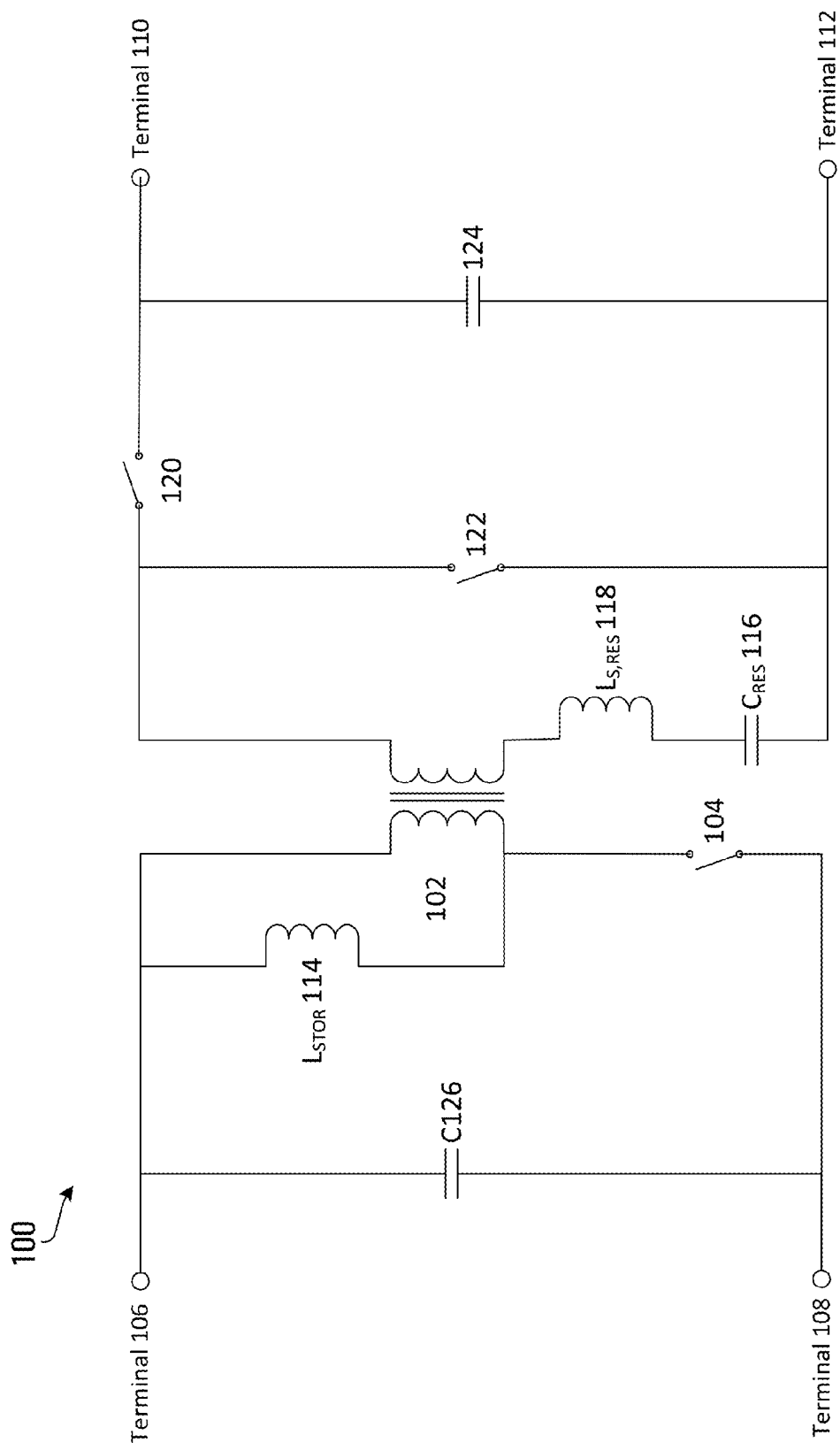


FIG. 1A

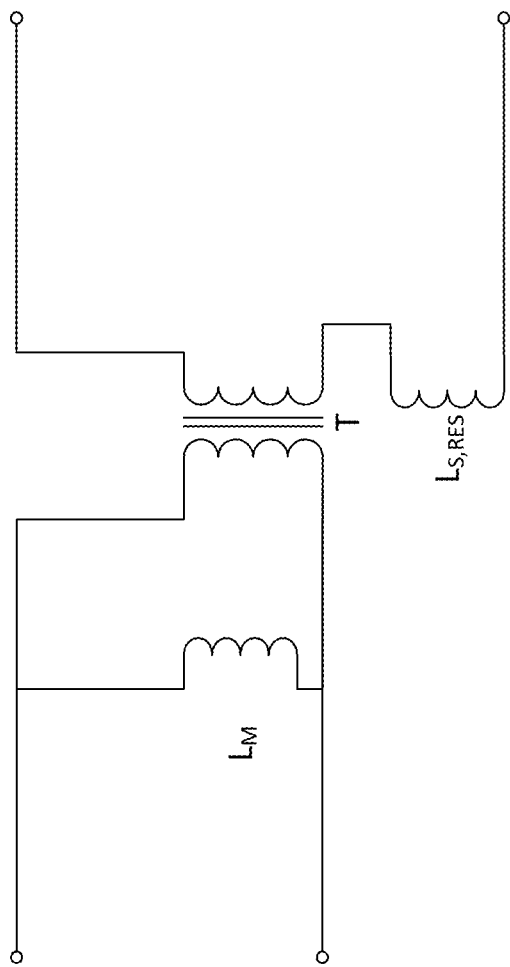


FIG. 1B

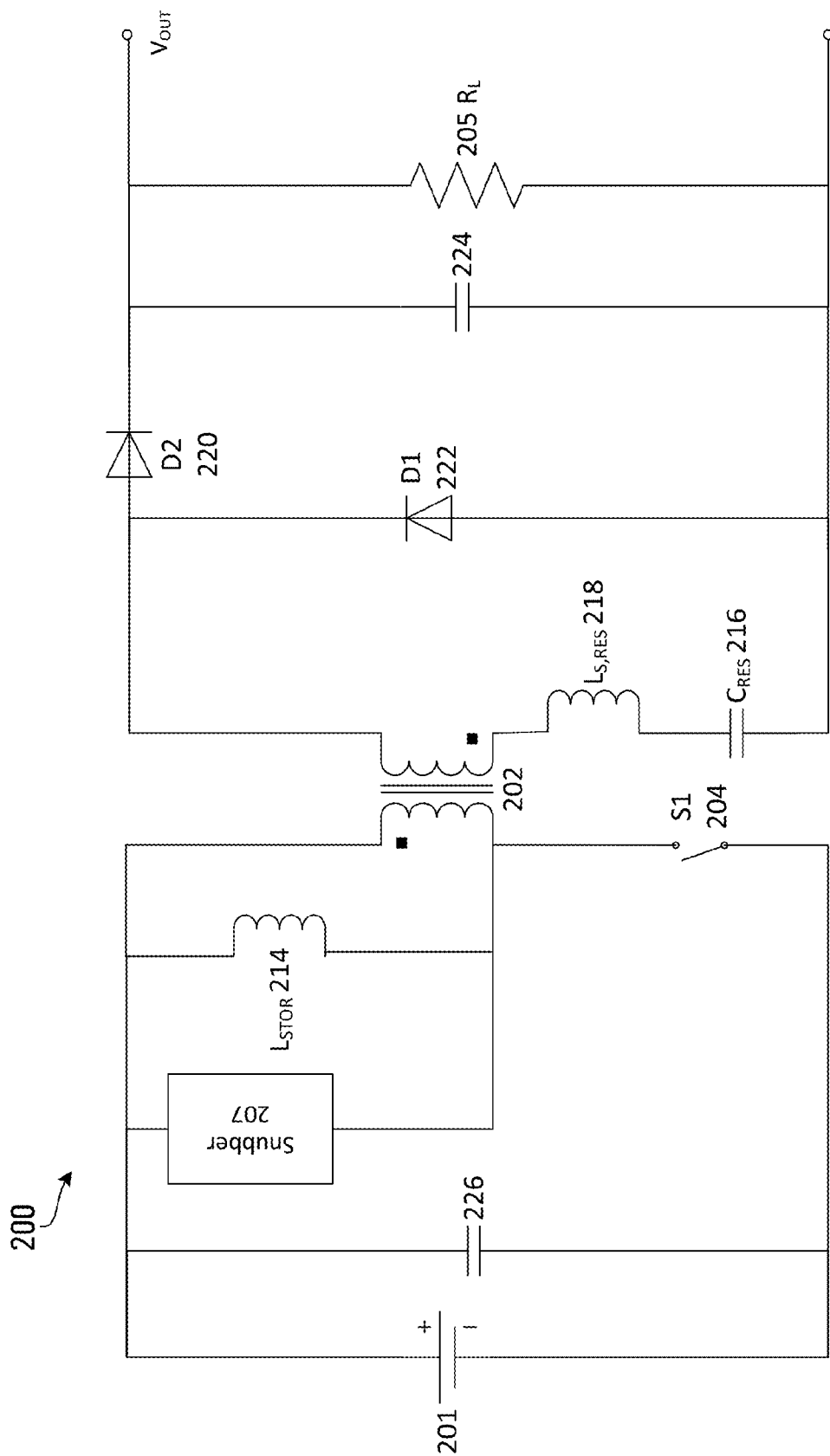


FIG. 2

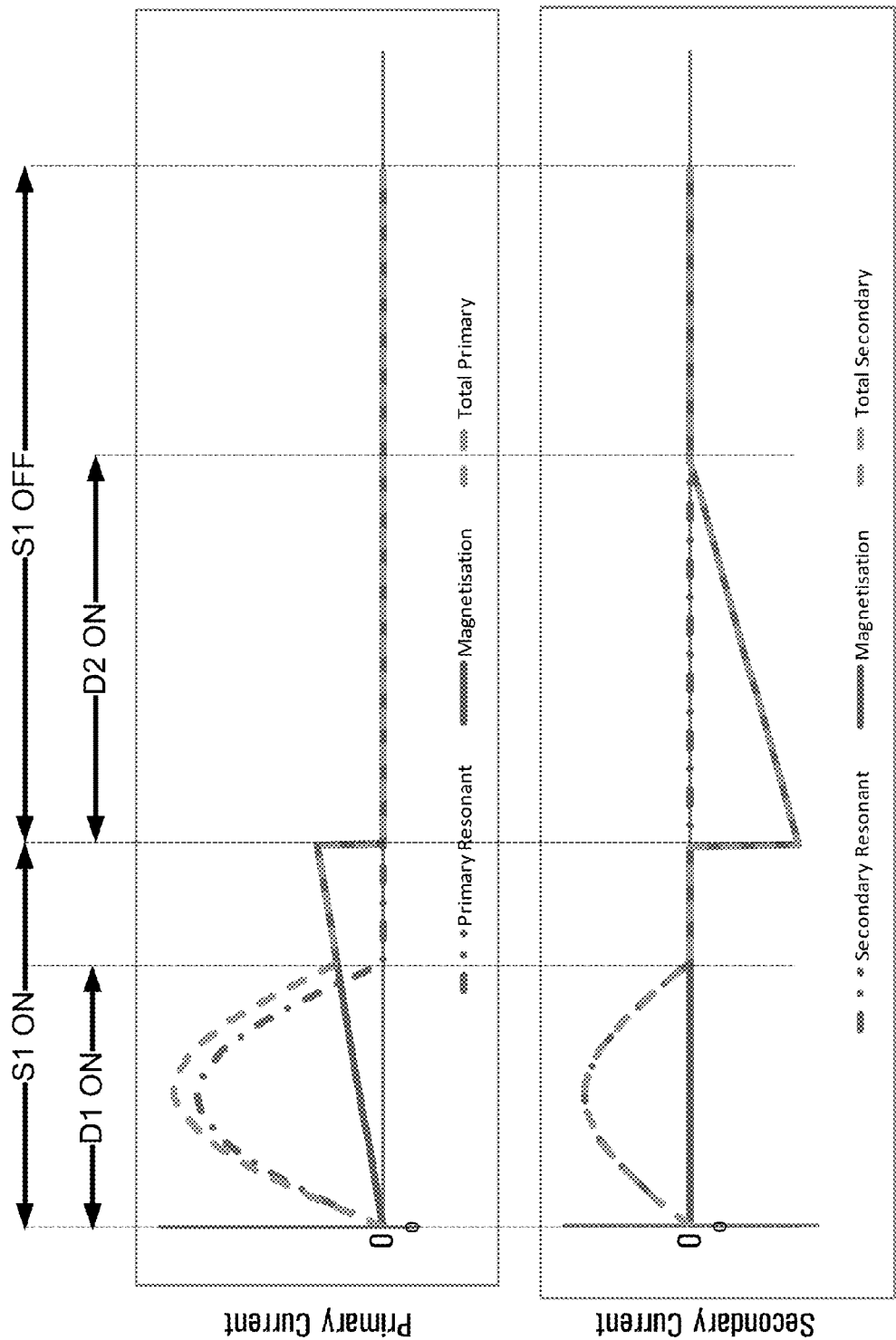


FIG. 3A

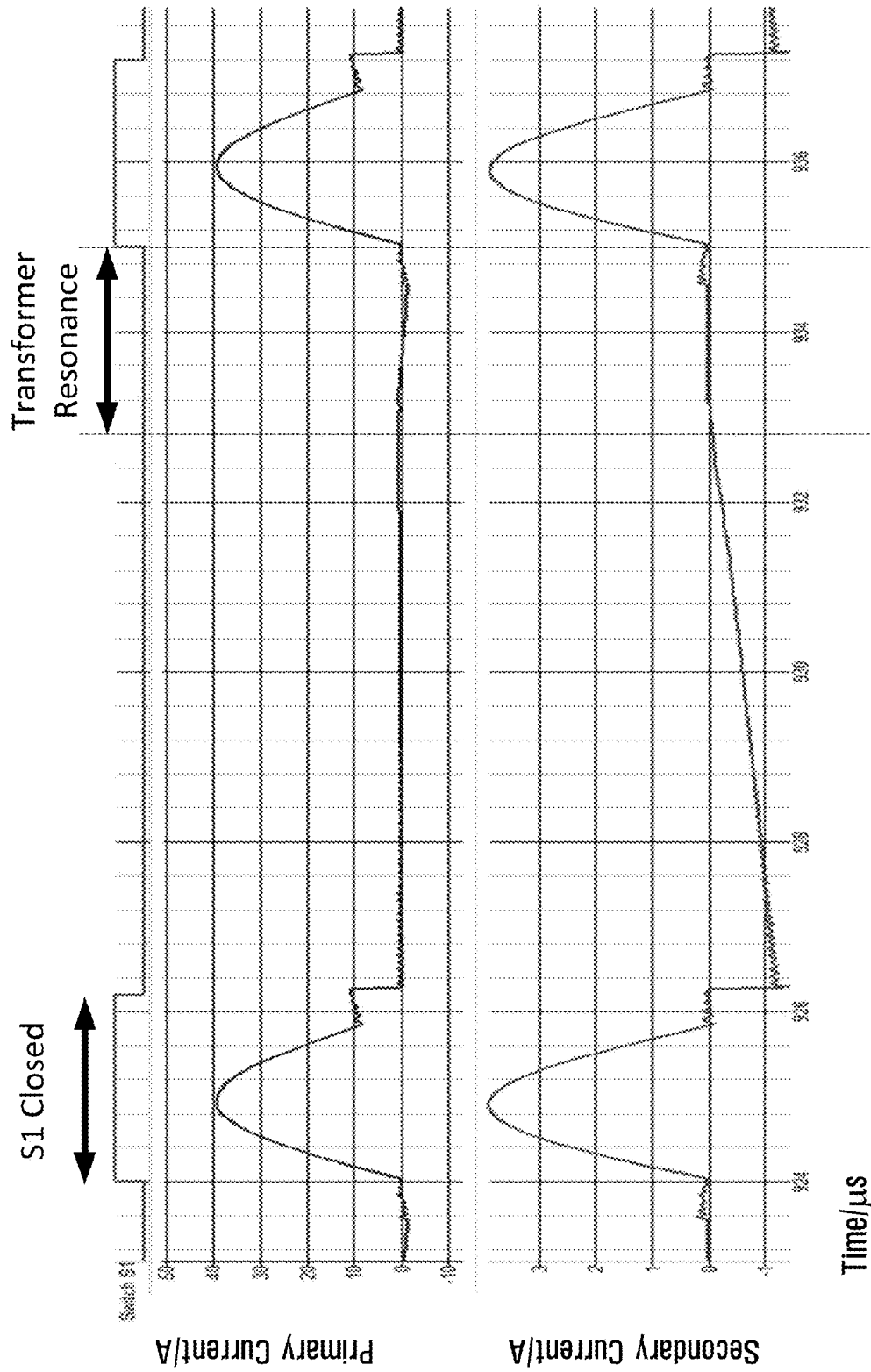


FIG. 3B

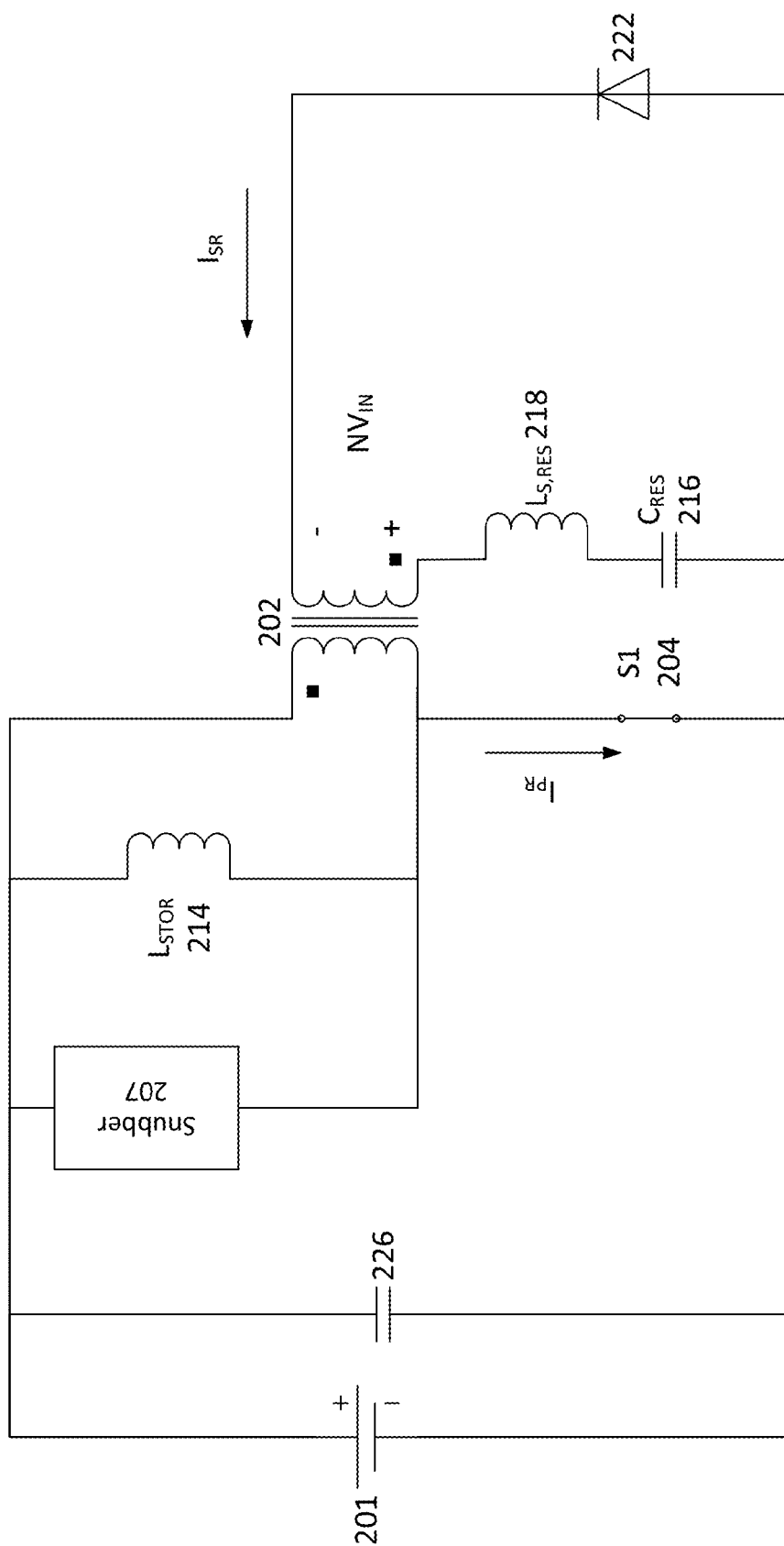


FIG. 4

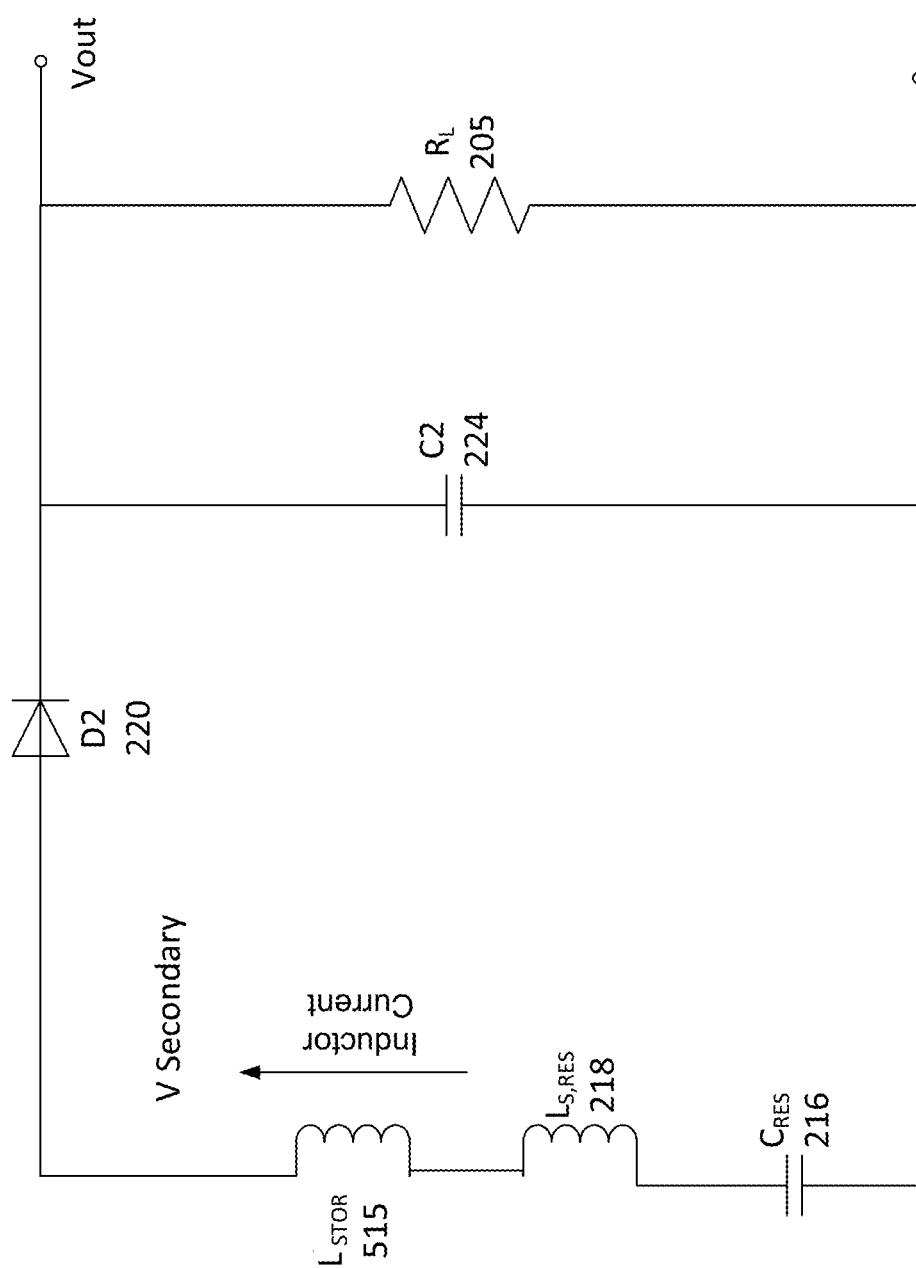


FIG. 5

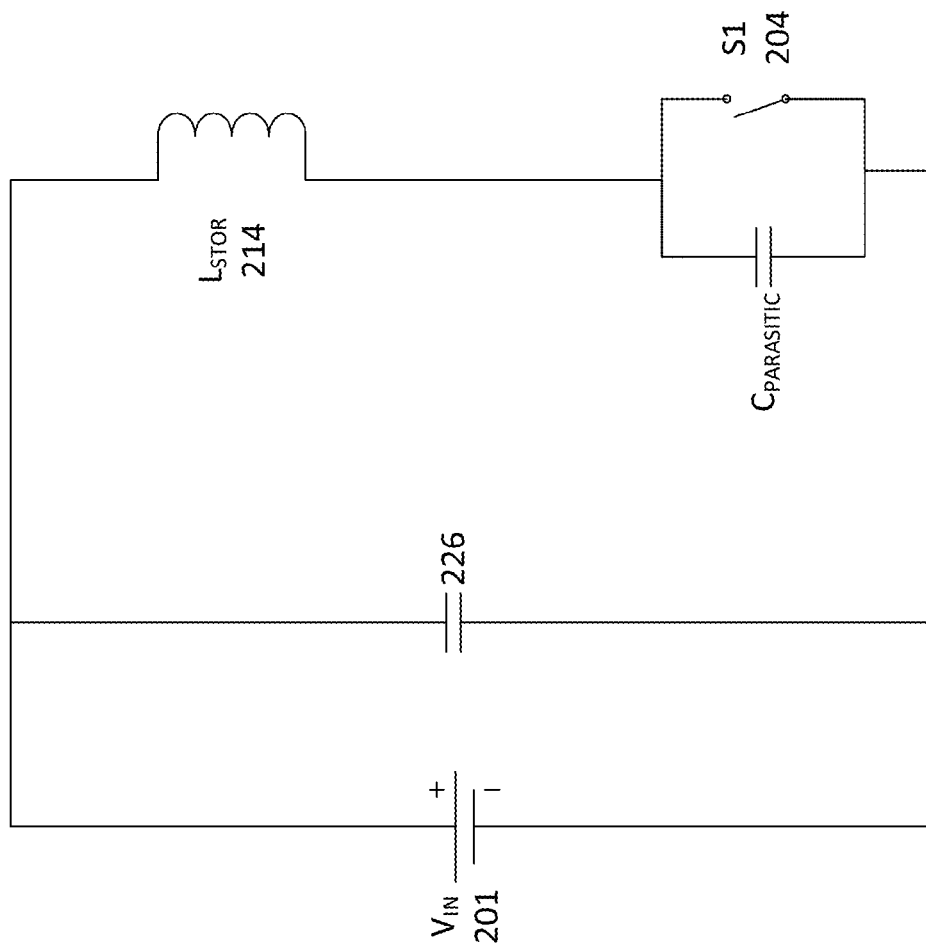


FIG. 6

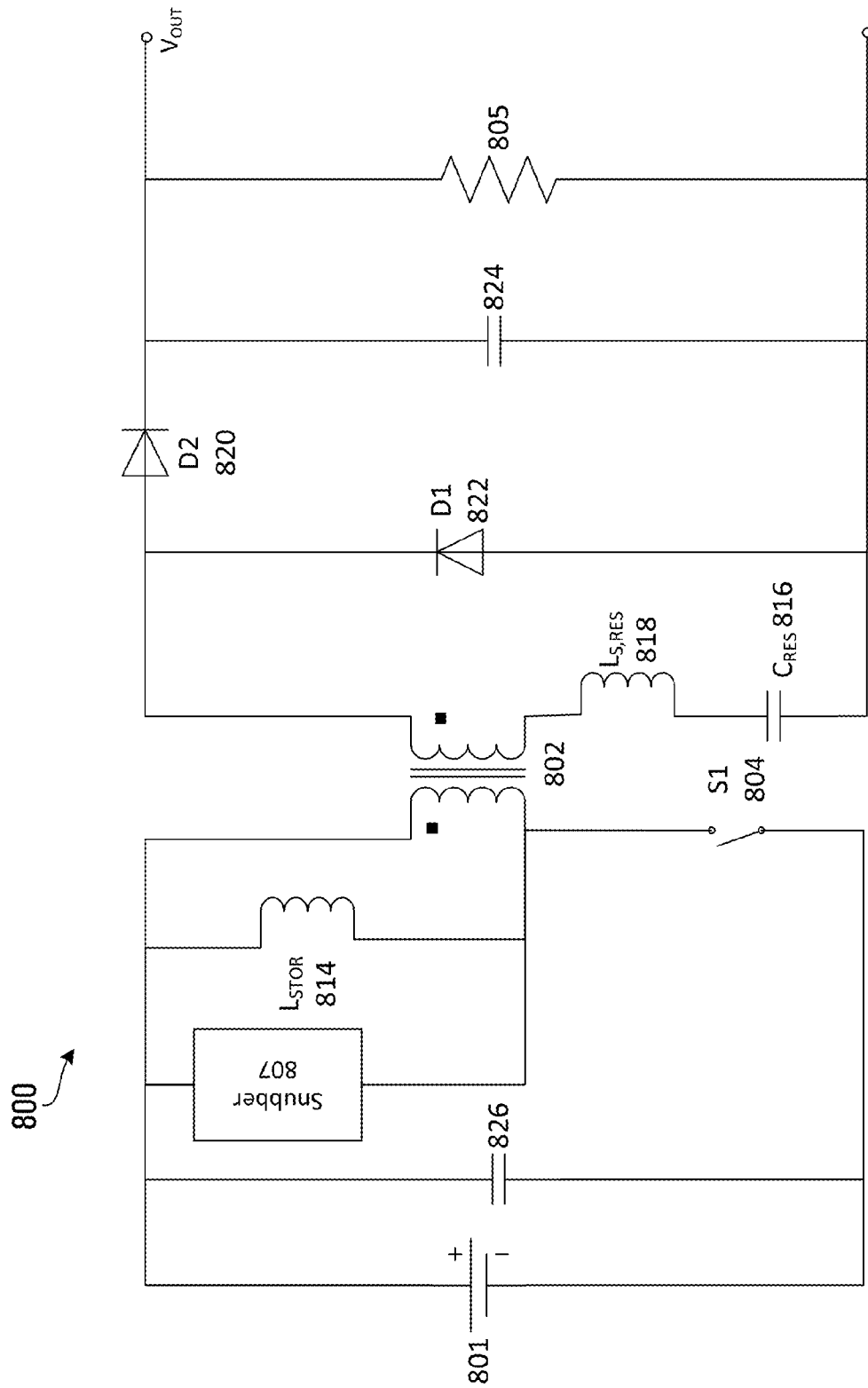


FIG. 7

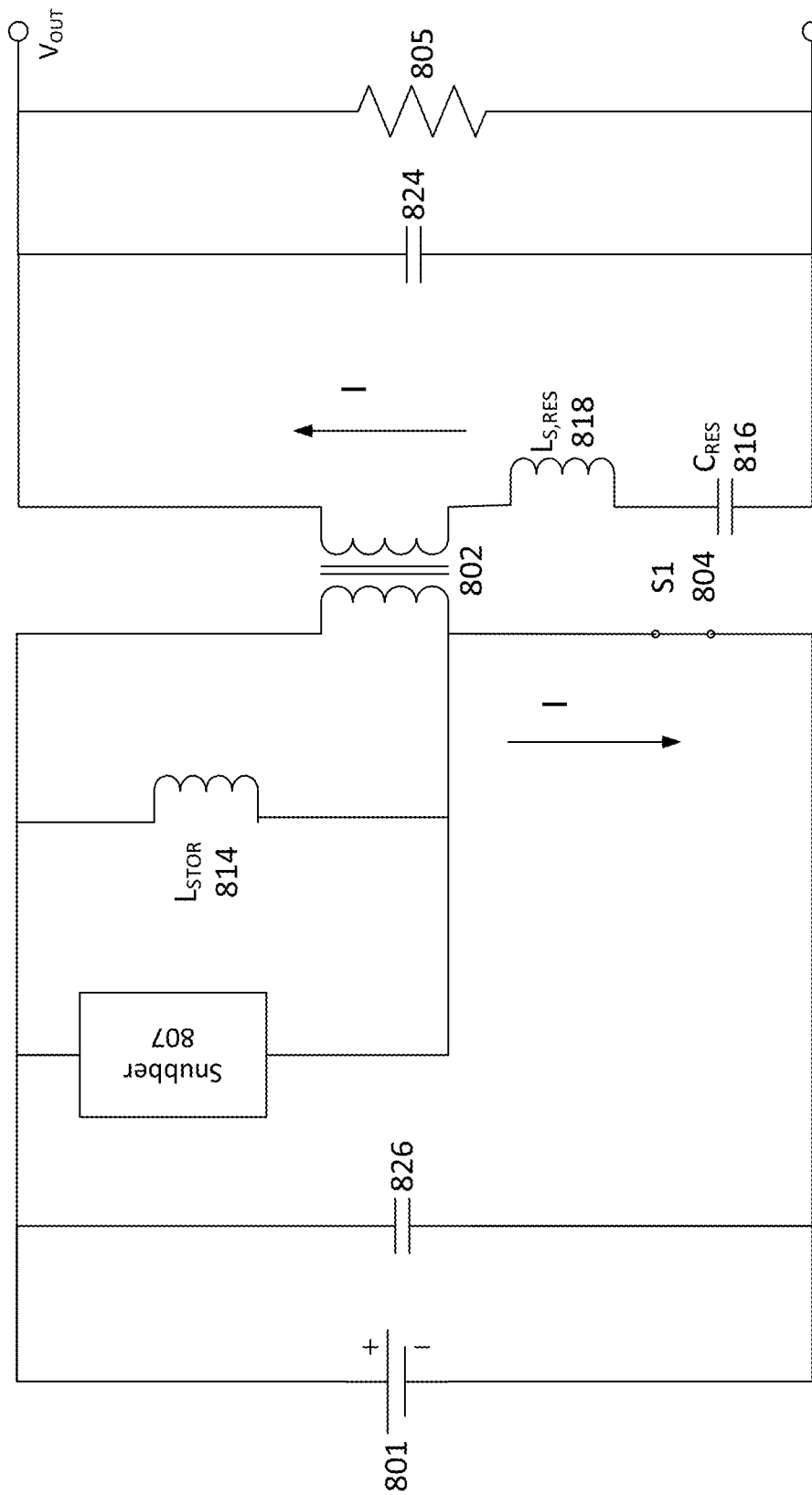


FIG. 8

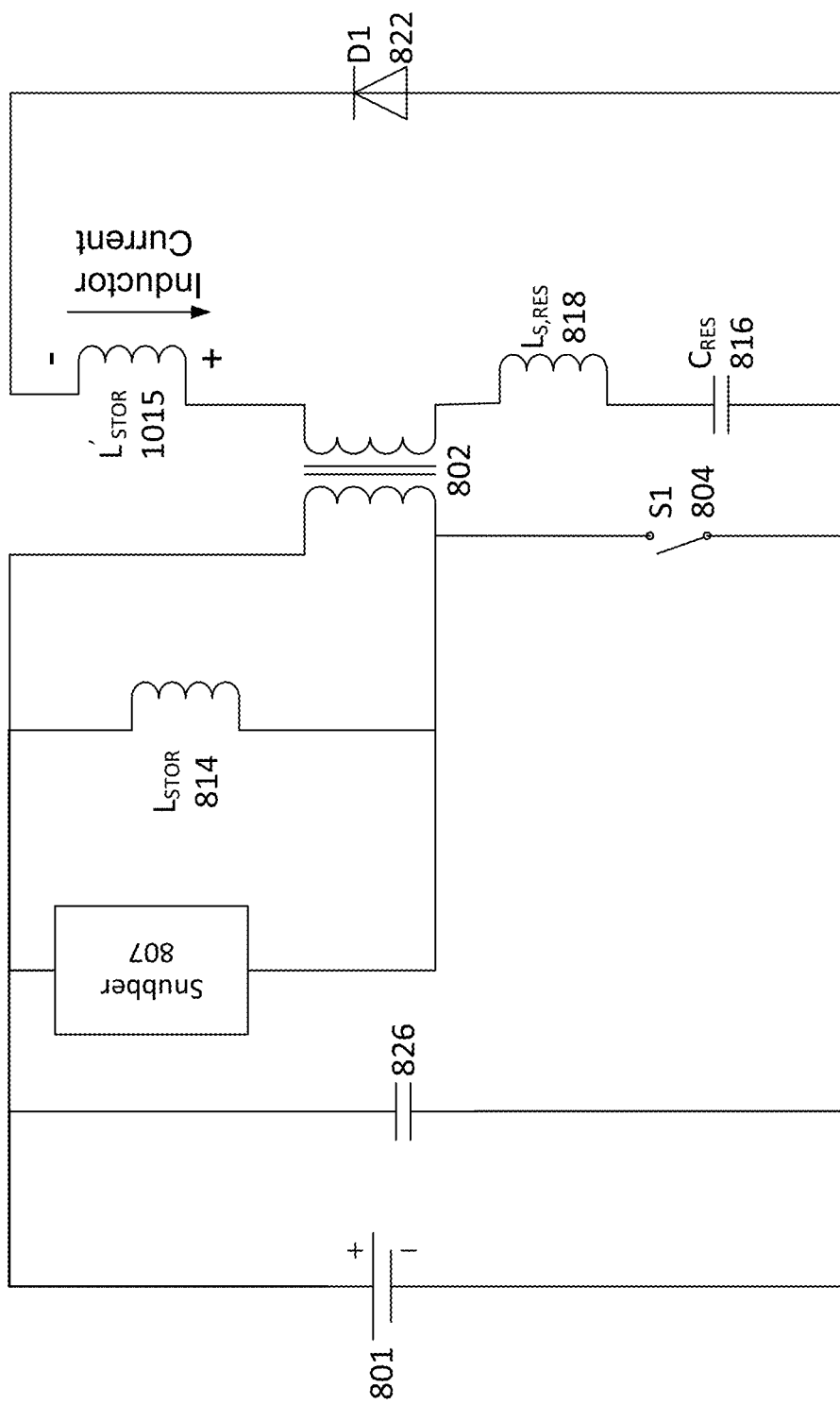


FIG. 9

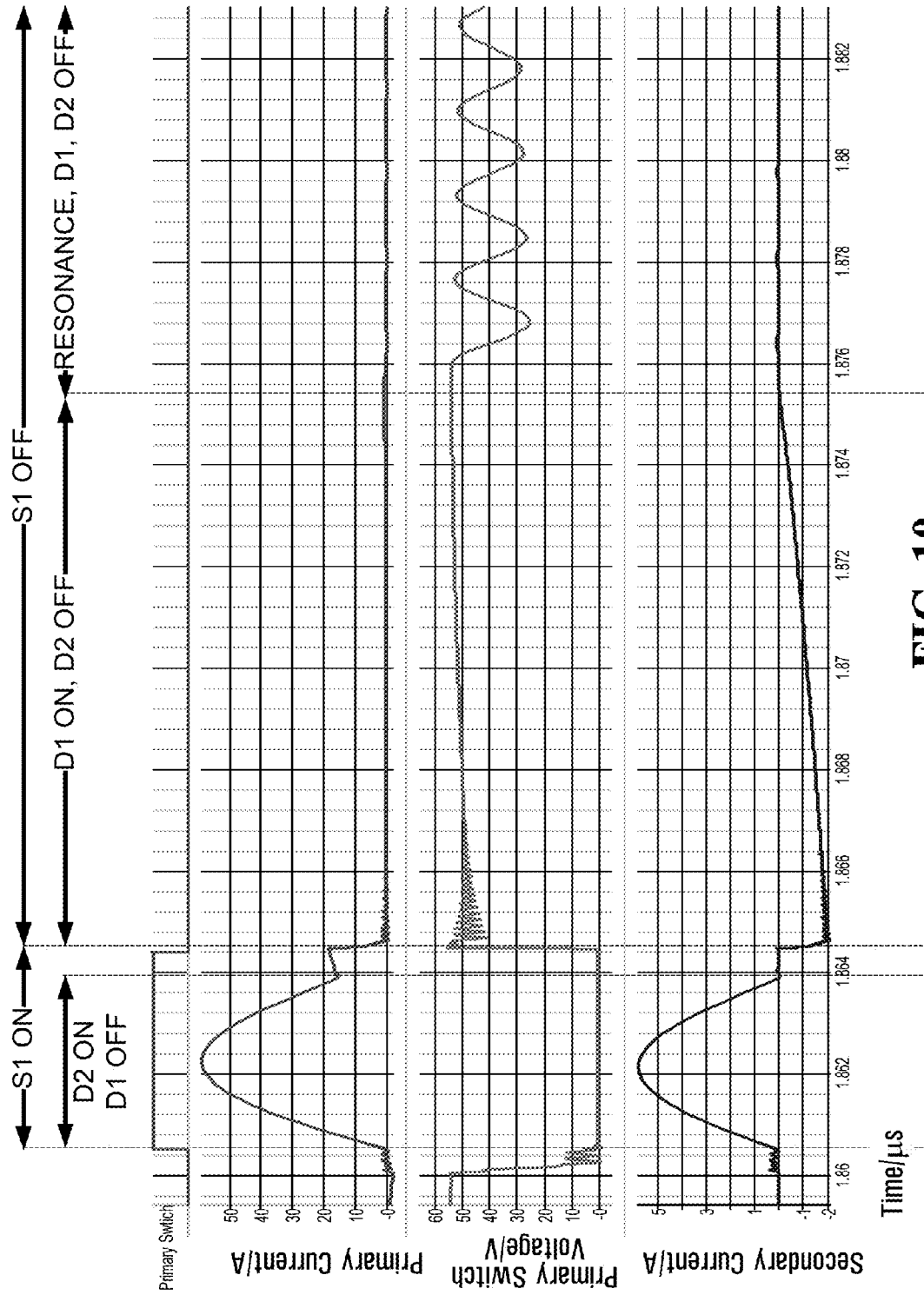


FIG. 10

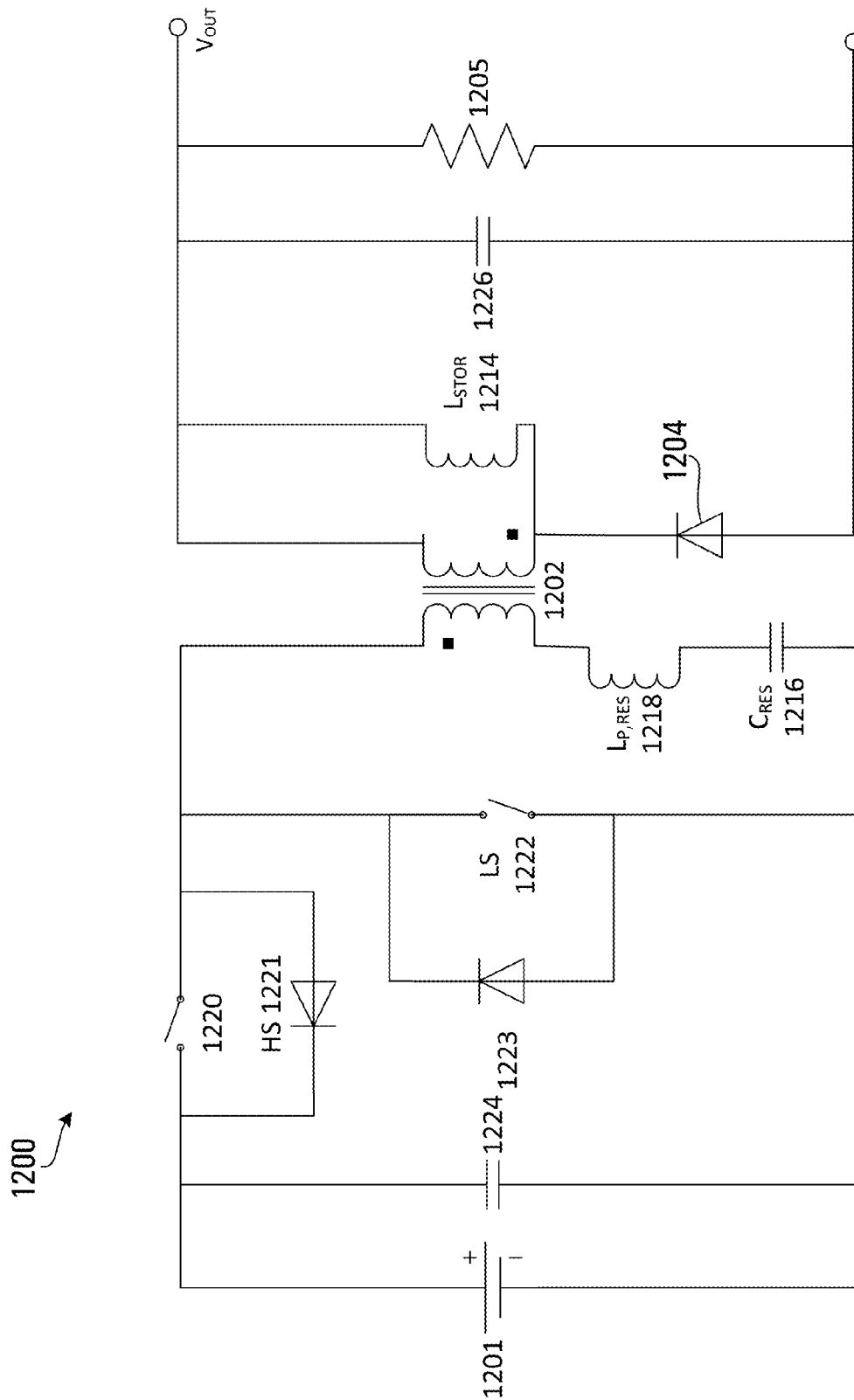


FIG. 11

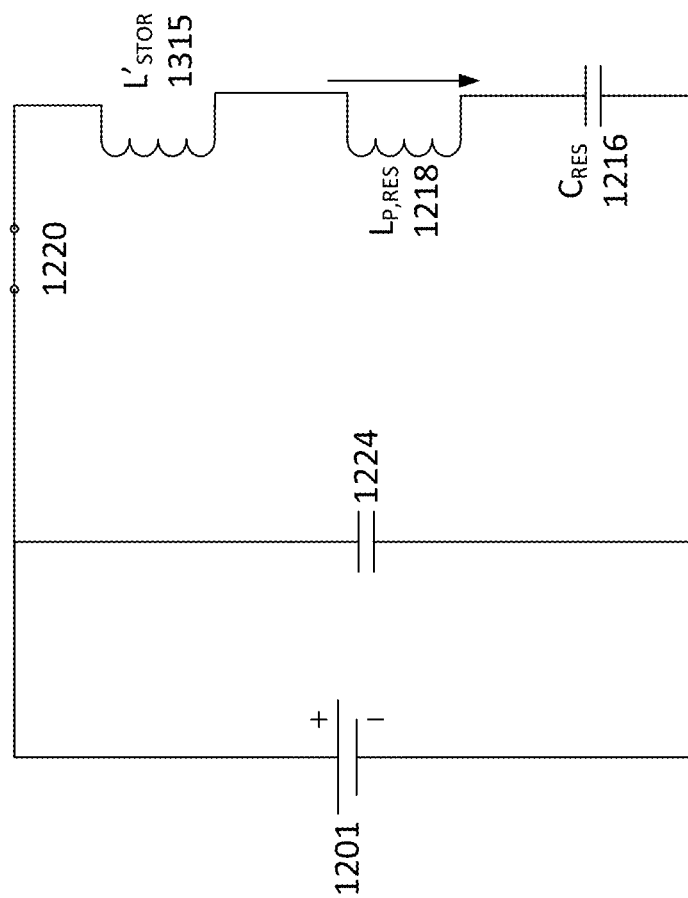


FIG. 12A

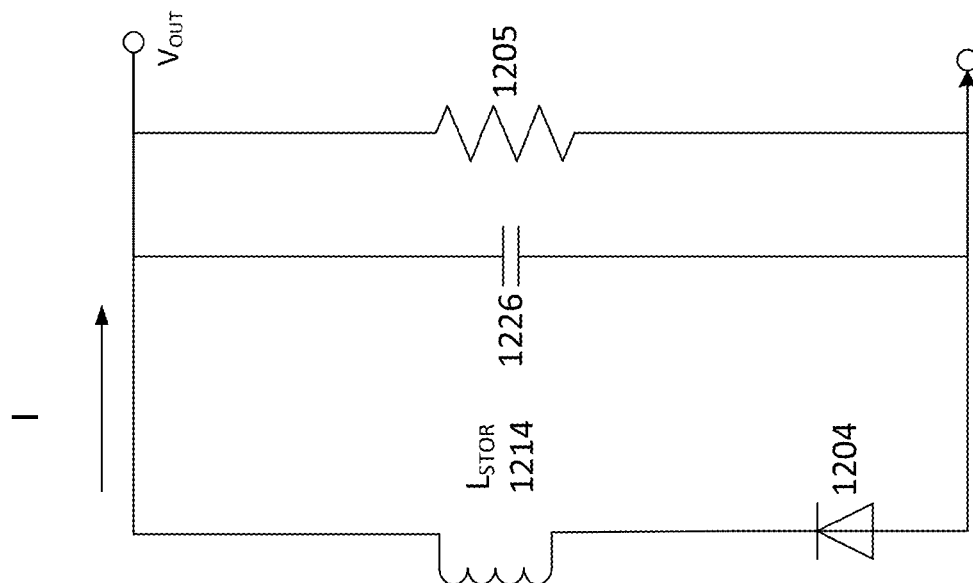


FIG. 12C

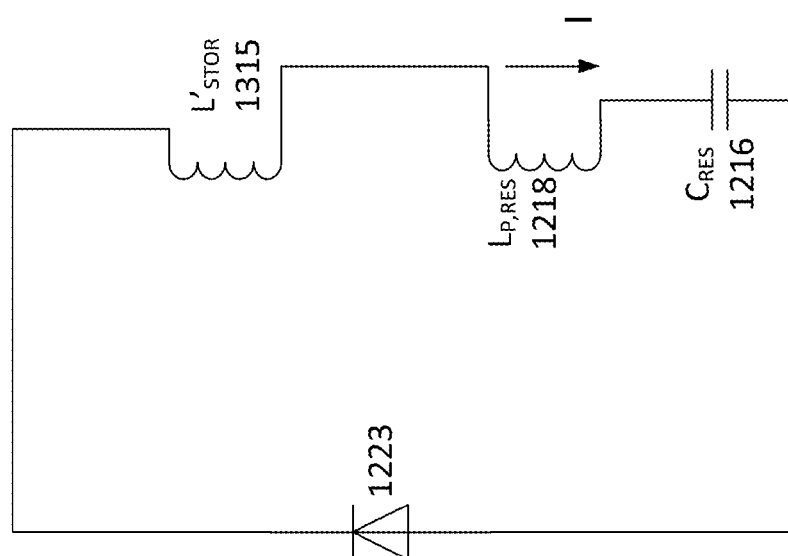


FIG. 12B

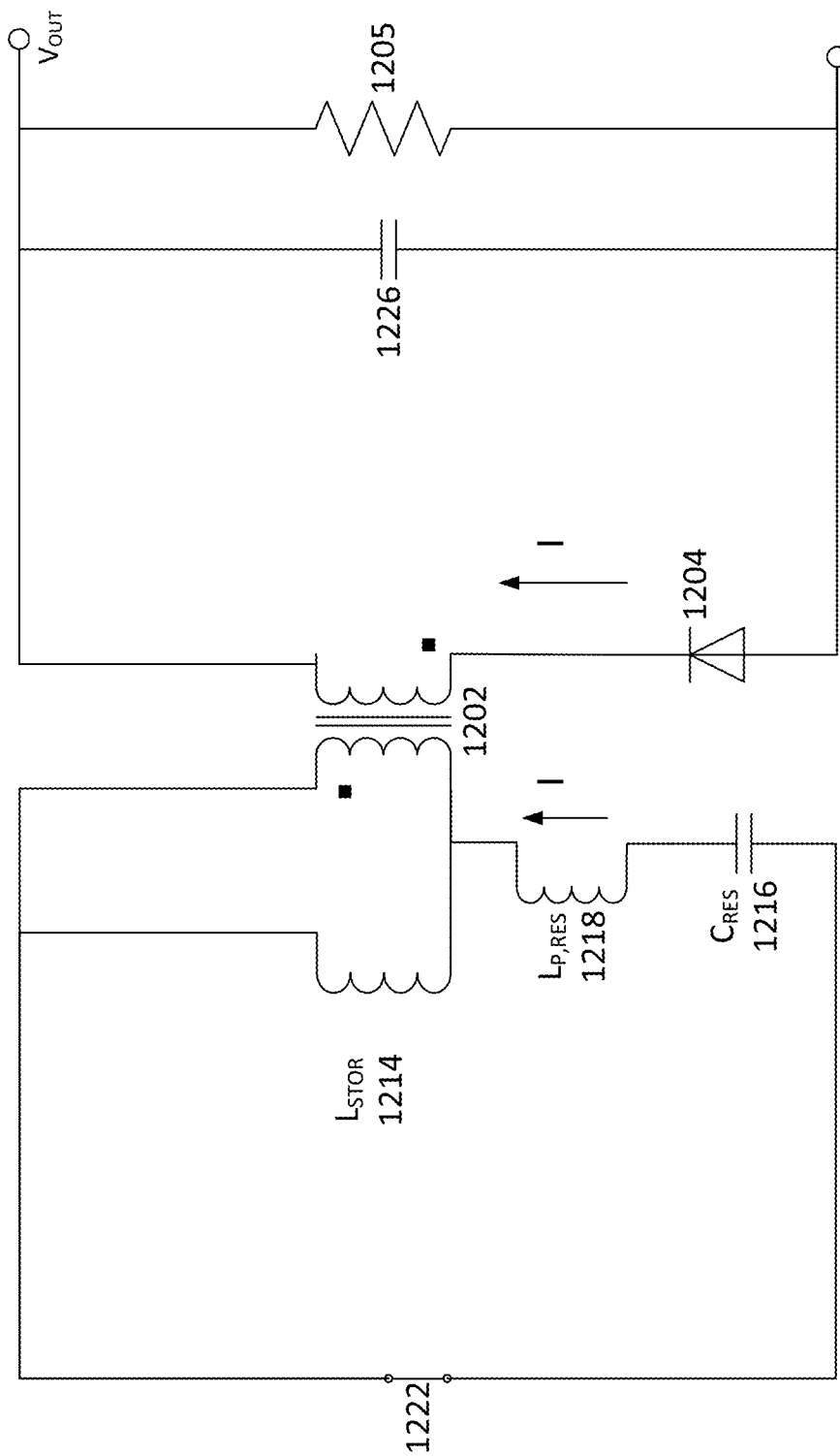


FIG. 12D

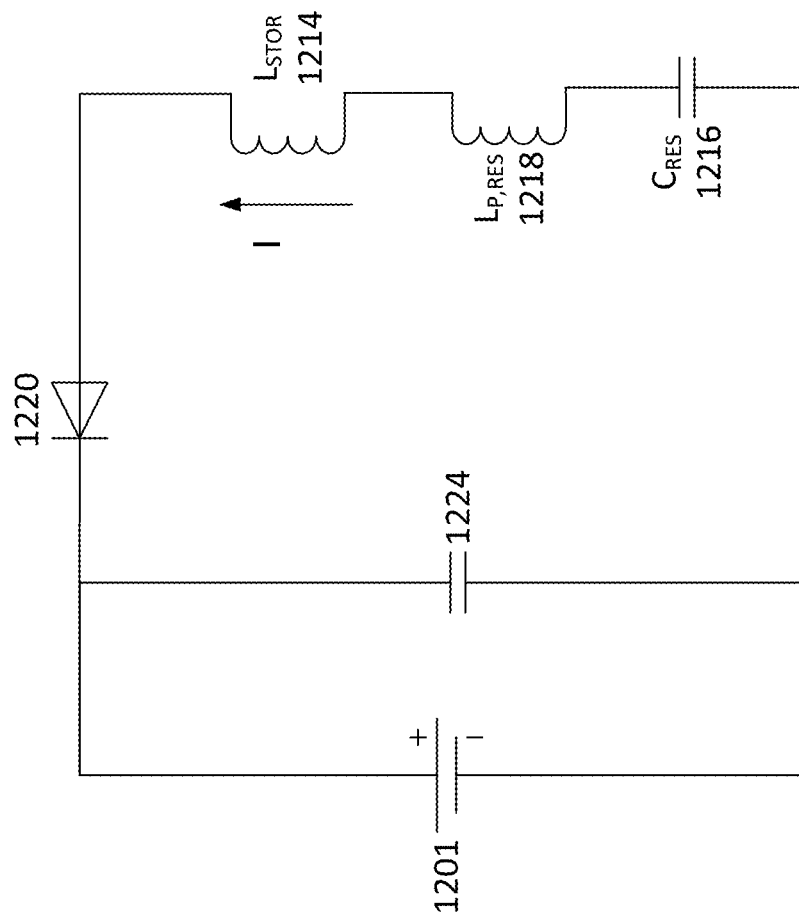


FIG. 12E

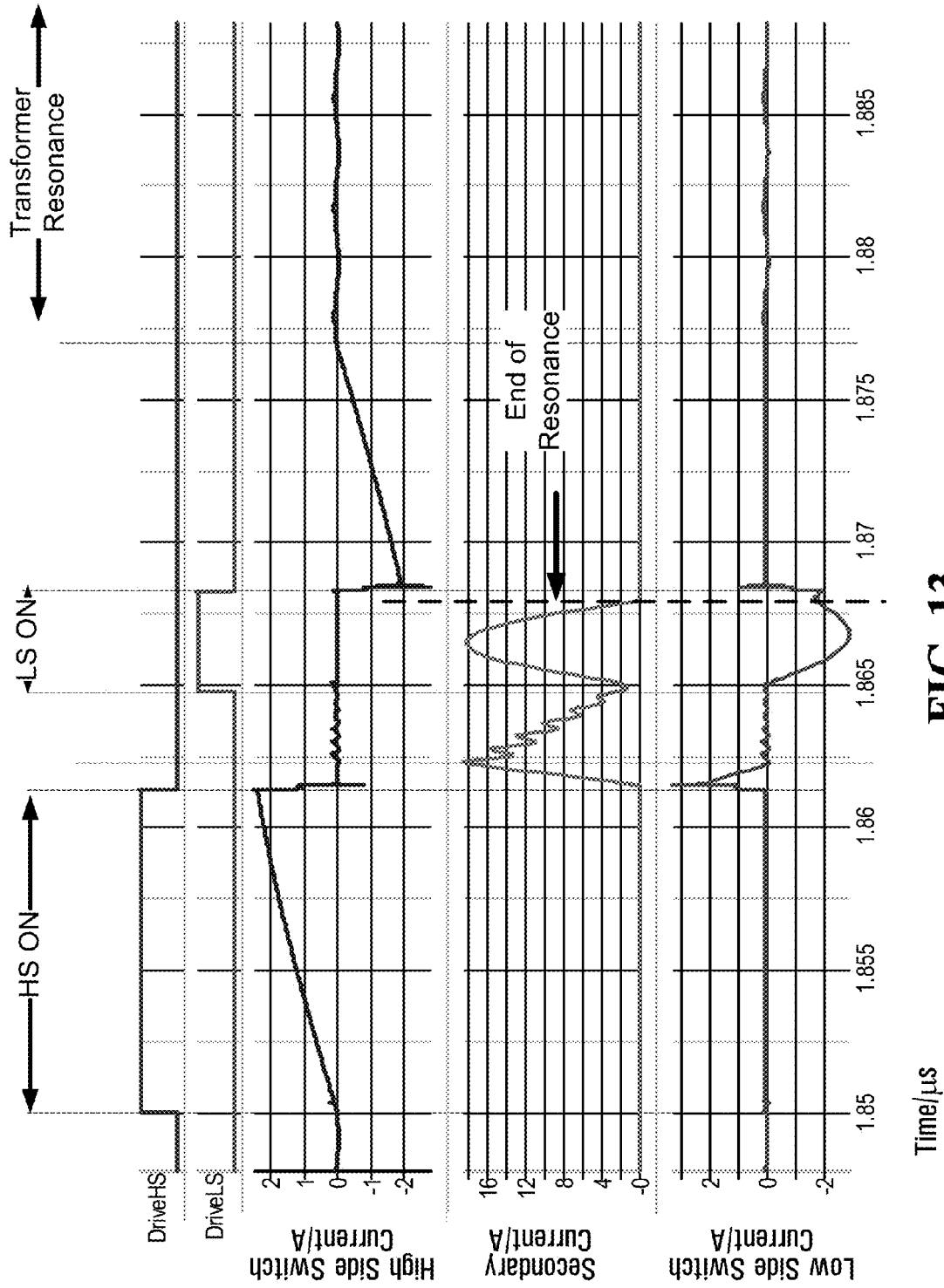


FIG. 13

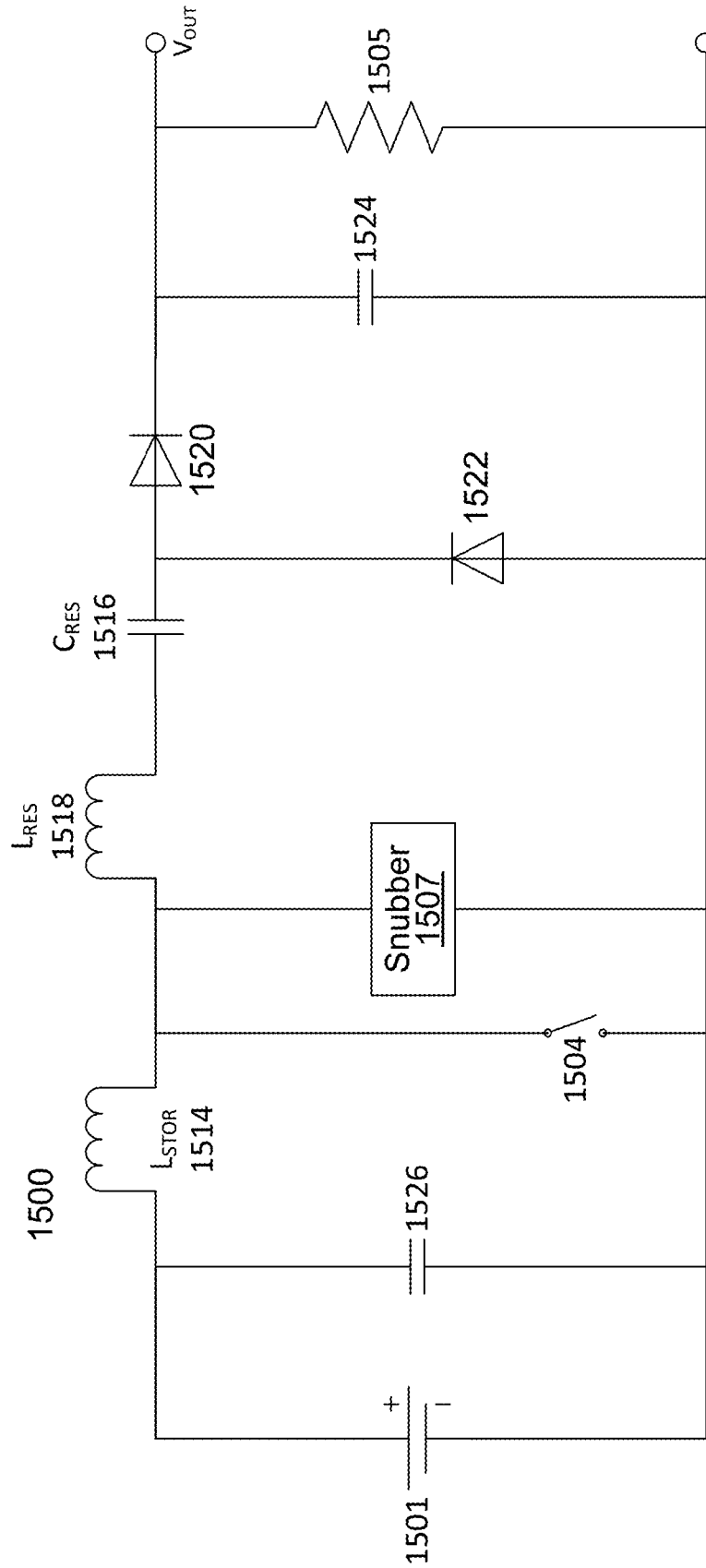


FIG. 14

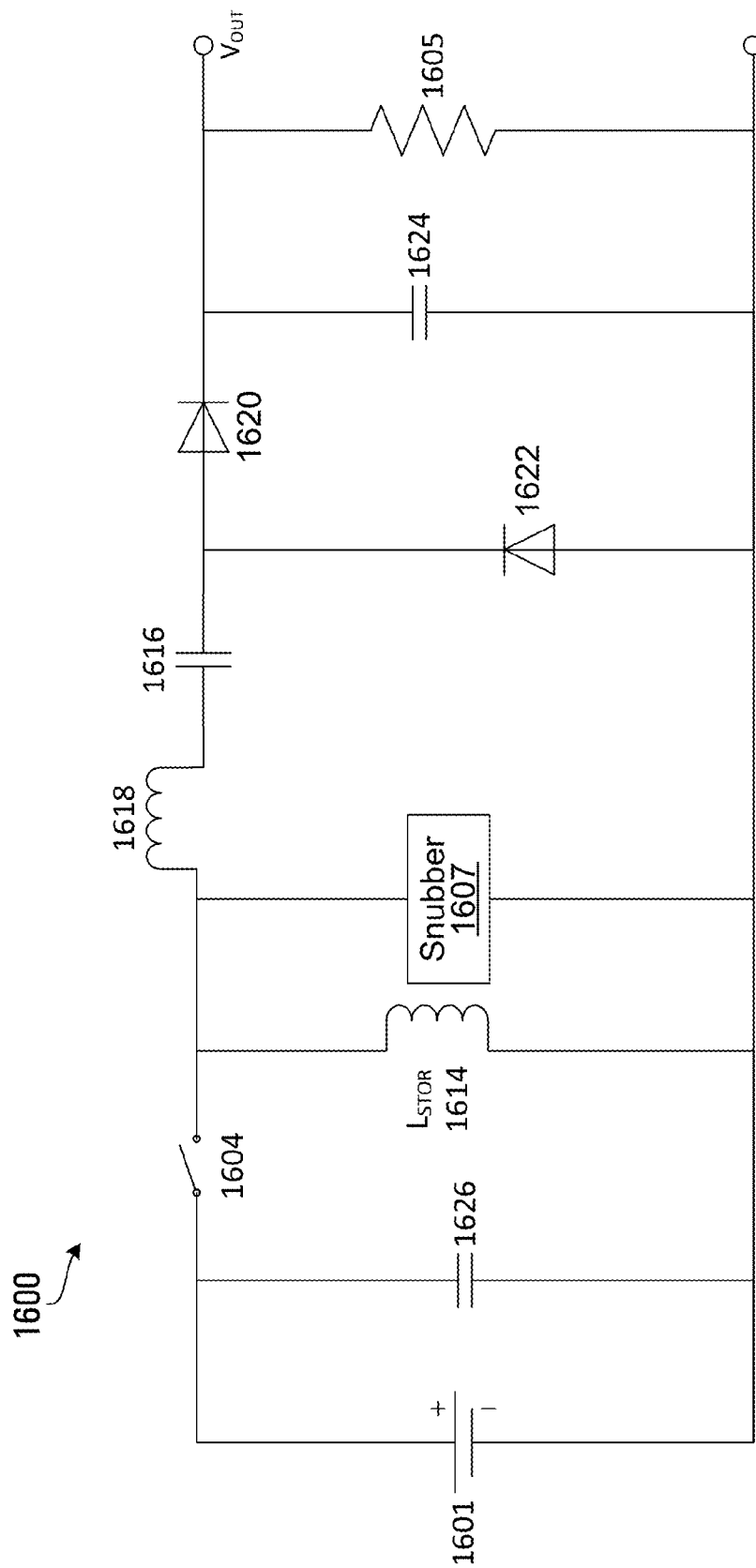


FIG. 15

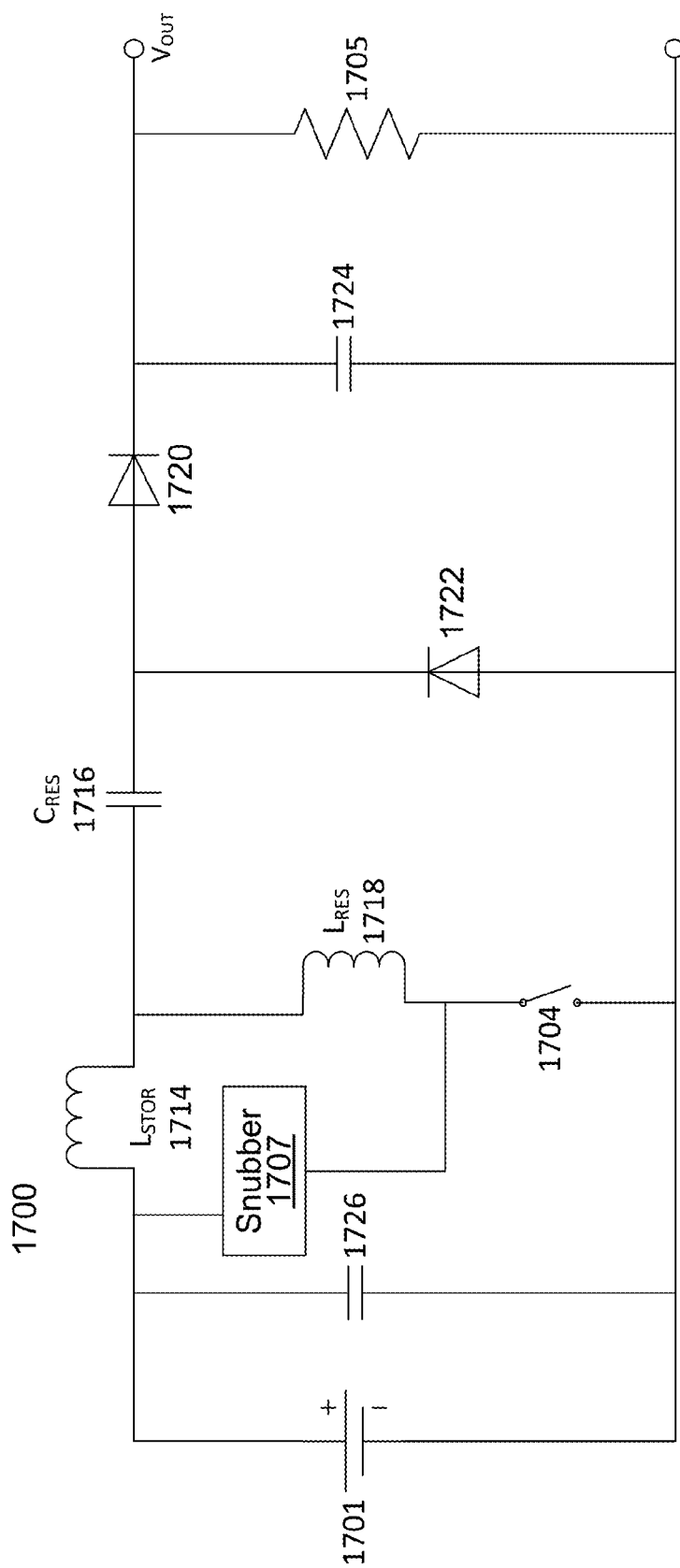


FIG. 16

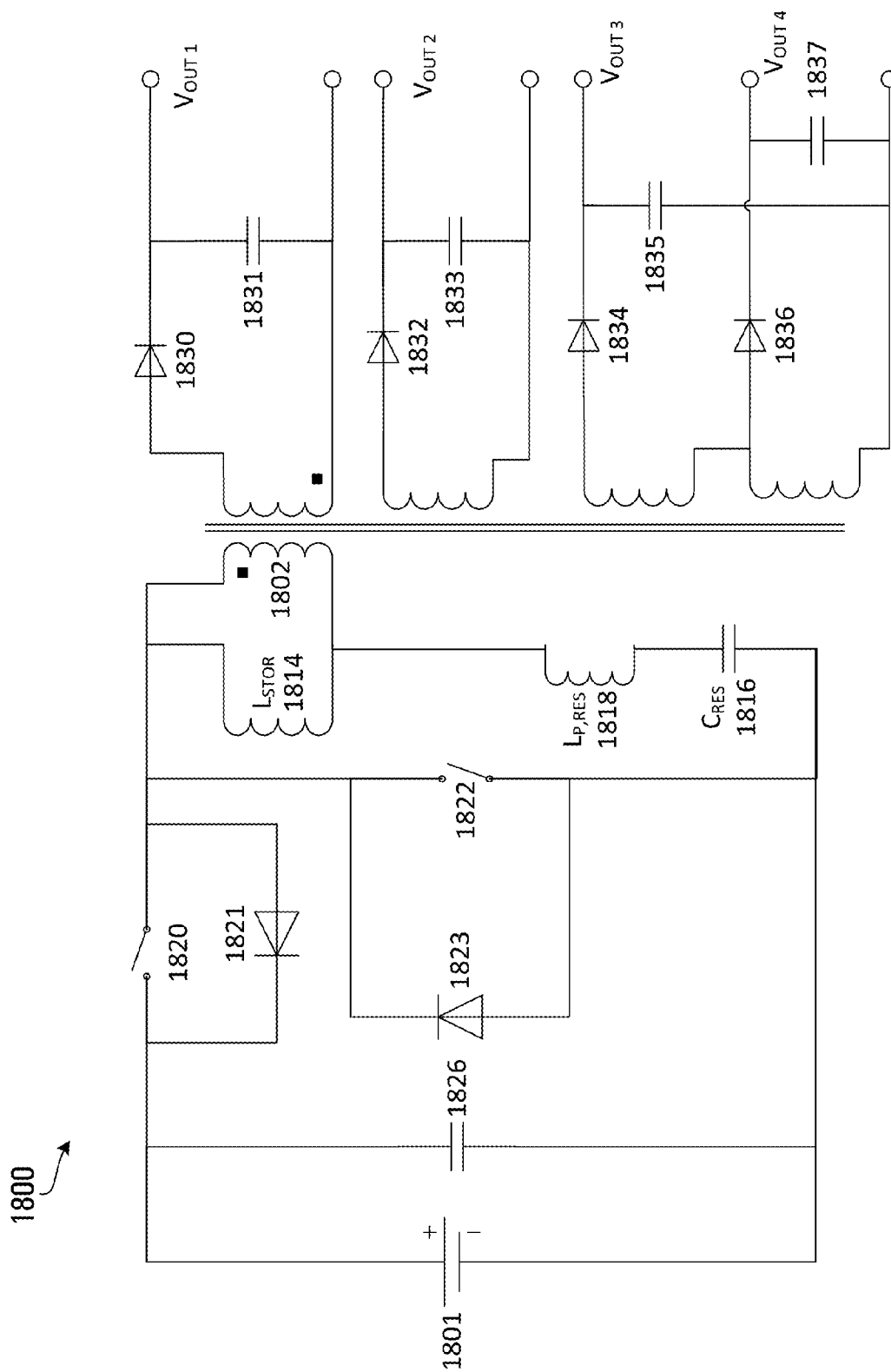


FIG. 17

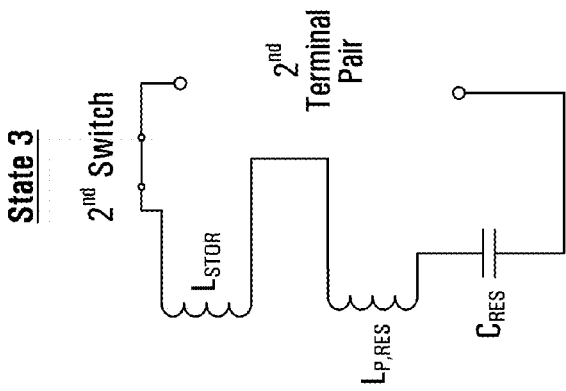


FIG. 20

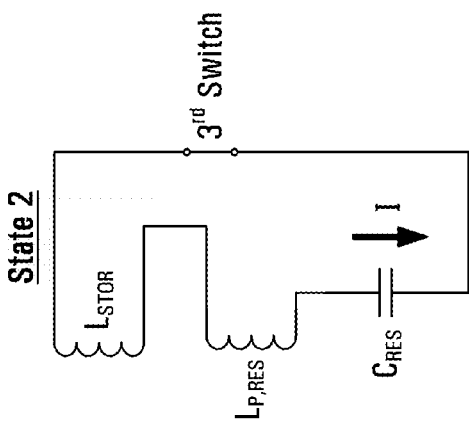


FIG. 19

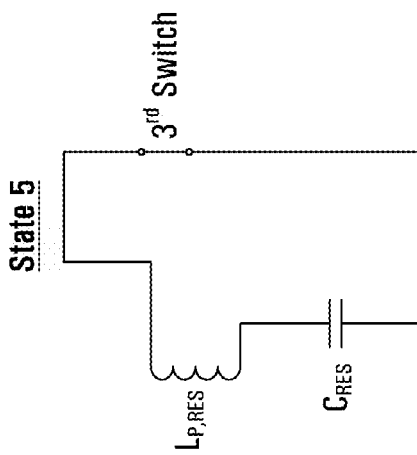


FIG. 22

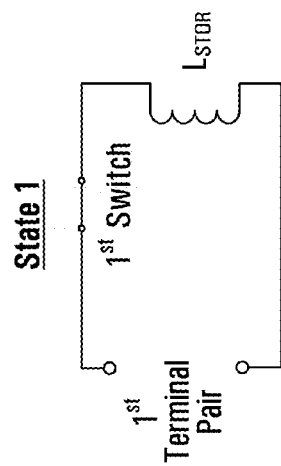


FIG. 18

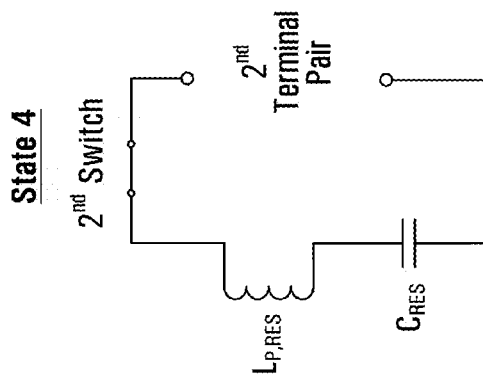


FIG. 21

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FORWARD BOOST POWER CONVERTERS AND METHODS

CROSS-REFERENCE TO RELATED APPLICATION

This application claims the benefit of U.S. Provisional Patent Application Ser. No. 61/711,529, filed on Oct. 9, 2012, entitled "FORWARD BOOST POWER CONVERTERS AND METHODS", the entire contents of which are incorporated herein by reference.

FIELD OF THE INVENTION

This invention relates generally to switching power converters and more particularly to frequency modulated power converters with resonant topologies. Power converters are alternatively referred to as DC (Direct Current) to DC converters, DC to AC (Alternating Current) converters, AC to AC converters, AC to DC converters, DC or DC to DC regulators, AC or AC to DC regulators, switching mode power converters, switch mode regulators or power supplies, etc. The term "power converter" or "converter" is used herein to include all such terms.

BACKGROUND

A power converter is a circuit which converts current at one voltage to current at another voltage. Switching converters use one or more switches to effect this conversion. Converters that control the output voltage by using a constant "on" time for the switch(es) and varying the switching frequency are called frequency modulated converters. Resonant converters are converters that use inductive and capacitive (LC) reactive elements in resonance to transfer power from an input source to the output.

Design objectives of power converters include: maximizing power conversion efficiency from input to output; minimizing component count and cost; and minimizing stress on the various converter components to maximize reliability. Another important design objective is a monotonic power transfer function. The transfer function relates the output power of the converter to its control variable. For frequency modulated converters, the control variable is the switching frequency.

SUMMARY

According to an aspect of the present disclosure, a switching mode power converter coupled between a first terminal pair and a second terminal pair includes: a first inductance coupled to a first switch in a first circuit path across said first terminal pair, said first switch coupling said first inductance across said first terminal pair during a conduction period of said first switch; a capacitance coupled to a second inductance in a second circuit path, and to said first inductance in a third circuit path; a second switch and a third switch, said second switch completing a circuit between said second terminal pair and one of: said second circuit path and said third circuit path during a conduction period of said second switch, said third switch completing the other of: said second circuit path and said third circuit path during a conduction period of said third switch.

Said first terminal pair and said second terminal pair may share a common terminal.

In some embodiments, said first terminal pair is an input terminal pair and said second terminal pair is an output ter-

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minal pair, with said capacitance transferring energy to said output terminal pair during said conduction period of said second switch.

Said first inductance may be coupled to said capacitance and transfer energy to said output terminal pair during said conduction period of said second switch.

Said first inductance could be coupled to said capacitance and transfer energy to said capacitance during said conduction period of said third switch.

In some embodiments, said first terminal pair is an output terminal pair and said second terminal pair is an input terminal pair, with said first inductance transferring energy to said output terminal pair during said conduction period of said first switch.

Said first inductance may be coupled to said capacitance and store energy from said input terminal pair during said conduction period of said second switch.

Said first inductance and said capacitance could transfer energy to said output terminal pair during said conduction period of said first switch.

In some embodiments, said first inductance is coupled to said capacitance through a transformer. Said first inductance could then include magnetisation inductance of said transformer. Said second inductance could include leakage inductance of said transformer.

Said transformer could include a first winding coupled across said first inductance and a second winding coupled in said second circuit path with said capacitance and said second inductance. Said transformer could further include a third winding.

If said transformer further includes a third winding, said switching mode power converter could further include a fourth switch coupled to said third winding and a reflected inductance of said transformer magnetisation inductance in a fourth circuit path across a third terminal pair, said fourth switch having said conduction period of said third switch and coupling said fourth circuit path across said third terminal pair during said conduction period of said third and fourth switches.

In another embodiment, said transformer further includes a third winding, and said switching mode power converter further includes: a fourth switch coupled to said third winding and a first reflected inductance of said transformer magnetisation inductance in a fourth circuit path across a third terminal pair, said fourth switch having said conduction period of said third switch and coupling said fourth circuit path across said third terminal pair during said conduction period of said third and fourth switches; a fifth switch coupled to a portion of said third winding and a second reflected inductance of said transformer magnetisation inductance in a fifth circuit path across a fourth terminal pair, said fifth switch having said conduction period of said third switch and coupling said fifth circuit path across said fourth terminal pair during said conduction period of said third and fifth switches.

The switching mode power converter could include one or more of: a capacitance coupled across said first terminal pair; and a capacitance coupled across said second terminal pair.

In some embodiments, said first inductance resonates during a non-conduction period of said first switch, said second switch and said third switch to produce a time varying voltage across said first switch.

Said first switch could enter said conduction period while said voltage is at an oscillatory minimum.

A method of operation of a switching mode power converter is also provided. The switching mode power converter is coupled between a first terminal pair and a second terminal pair, the converter including: a first inductance coupled to a

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first switch in a first circuit path across said first terminal pair; a capacitance coupled to a second inductance in a second circuit path, and to said first inductance in a third current circuit path; a second switch and a third switch. The method involves: closing of said first switch to couple said first inductance across said first terminal pair during a conduction period of said first switch; closing of said second switch to complete a circuit between said second circuit path or said third circuit path and said second terminal pair during a conduction period of said second switch so as to transfer energy into said second terminal pair; closing of said third switch to complete said second circuit path or said third circuit path during a conduction period of said third switch so as to transfer energy into said capacitance.

According to another aspect of the present disclosure, an apparatus includes: a switching mode power converter coupled between a first terminal pair and a second terminal pair, the converter including: a first inductance coupled to a first switch in a first circuit path across said first terminal pair; a capacitance coupled to a second inductance in a second circuit path, and to said first inductance in a third current circuit path; a second switch and a third switch; means for closing said first switch to couple said first inductance across said first terminal pair during a conduction period of said first switch; means for closing said second switch to complete a circuit between said second circuit path or said third circuit path and said second terminal pair during a conduction period of said second switch so as to transfer energy into said second terminal pair; means for closing said third switch to complete said second circuit path or said third circuit path during a conduction period of said third switch so as to transfer energy into said capacitance.

Another aspect of the present disclosure provides a method of energy transfer in a switching mode power converter. The method involves transfer of energy between an inductance and a first terminal pair by a substantially linearly varying current; and one of: transfer of energy resonantly between a capacitance and a second terminal pair by a substantially half sinusoidal current pulse and transfer of energy between said inductance and said capacitance by a substantially linearly varying current; and resonance of said capacitance for a half sinusoid cycle and transfer of energy between said inductance and said second terminal pair by a substantially linearly varying current.

The method could involve operation of said switching mode power converter in a forward direction wherein energy is transferred from said first terminal pair to said second terminal pair, and/or operation of said switching mode power converter in a reverse direction wherein energy is transferred from said second terminal pair to said first terminal pair.

Other aspects and features of embodiments of the present disclosure will become apparent to those ordinarily skilled in the art upon review of the following description.

BRIEF DESCRIPTION OF THE DRAWINGS

Examples of embodiments of the present disclosure will now be described in greater detail with reference to the accompanying drawings.

FIG. 1A is a schematic diagram illustrating an example forward boost converter.

FIG. 1B is a schematic diagram illustrating a transformer's parasitic inductances.

FIG. 2 is a schematic diagram illustrating an example forward boost converter using an anti-phase transformer.

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FIG. 3A is a current versus time graph illustrating resonant and storage current waveforms in the primary and secondary windings for the example forward boost converter **200** in FIG. 2.

FIG. 3B is a current versus time graph illustrating simulated primary and secondary current waveforms for the example forward boost converter **200** in FIG. 2.

FIG. 4 is a schematic diagram illustrating an equivalent circuit for the example forward boost converter **200** in FIG. 2 while the primary switch **204** is closed.

FIG. 5 is a schematic diagram illustrating an equivalent circuit for the example forward boost converter **200** in FIG. 2 when the primary switch **204** opens.

FIG. 6 is a schematic diagram illustrating an equivalent circuit for the example forward boost converter **200** in FIG. 2 during a transformer resonance period.

FIG. 7 is a schematic diagram illustrating an example forward boost converter circuit using a transformer with in-phase windings.

FIG. 8 is a schematic diagram illustrating an equivalent circuit for the example forward boost converter **800** in FIG. 7 while the primary switch **804** is closed.

FIG. 9 is a schematic diagram illustrating an equivalent circuit for the example forward boost converter **800** in FIG. 7 when the primary switch **804** opens.

FIG. 10 is a graph illustrating a simulated voltage waveform across the primary switch **804** and simulated primary and secondary current waveforms for the example forward boost converter **800** in FIG. 7.

FIG. 11 is a schematic diagram illustrating an example forward boost converter operated in the reverse direction to the example forward boost converter **200** in FIG. 2.

FIG. 12A is a schematic diagram illustrating an equivalent circuit for the example forward boost converter **1200** in FIG. 11 while high side switch **1220** is closed.

FIG. 12B is a schematic diagram illustrating an equivalent circuit for the input portion of the example forward boost converter **1200** in FIG. 11 when high side switch **1220** opens.

FIG. 12C is a schematic diagram illustrating an equivalent circuit for the output portion of the example forward boost converter **1200** in FIG. 11 when high side switch **1220** opens.

FIG. 12D is a schematic diagram illustrating an equivalent circuit for the input portion of the example forward boost converter **1200** in FIG. 11 when the low side switch **1222** is closed.

FIG. 12E is a schematic diagram illustrating an equivalent circuit for the input portion of the example forward boost converter **1200** in FIG. 11 when the low side switch opens.

FIG. 13 is a graph illustrating simulated high side and low side switch drive voltages and current waveforms for the high side switch **1220**, low side switch **1222** and secondary winding for the example forward boost converter **1200** in FIG. 11.

FIG. 14 is a schematic diagram illustrating another example forward boost converter.

FIG. 15 is a schematic diagram illustrating an alternate example of a forward boost converter.

FIG. 16 is a schematic diagram illustrating another example of a forward boost converter.

FIG. 17 is a schematic diagram illustrating an example forward boost converter with multiple secondary windings in reverse operation.

FIGS. 18 to 22 are schematic diagrams illustrating equivalent circuits for example forward boost converter operating states.

It should be appreciated that the contents of the drawings are intended solely for illustrative purposes, and that the

present disclosure is in no way limited to the particular example embodiments explicitly shown in the drawings and described herein.

DETAILED DESCRIPTION

For the purpose of promoting an understanding of the principles of the invention, reference will now be made to the embodiments illustrated in the drawings and specific language will be used to describe the same. It will nevertheless be understood that no limitation of the scope of the invention is thereby intended. Any alterations and further modifications in the described embodiments, and any further applications of the principles of the invention as described herein are contemplated as would normally occur to one skilled in the art to which the invention relates. One embodiment of the invention is shown in great detail, although it will be apparent to those skilled in the relevant art that some features that are not relevant to the present invention may not be shown for the sake of clarity.

FIG. 1A is a schematic diagram illustrating an example converter **100**. The converter **100** is an example of a converter topology which is referred to herein as “Forward Boost”. The primary winding of a transformer **102** is coupled at one end or terminal to a terminal **106** of a terminal pair **106, 108**. The other end or terminal of the primary winding is coupled to a primary switch **104**, which switchably couples the primary winding to the other terminal **108** of the terminal pair **106, 108**. Transformer **102** provides energy storage, and optionally voltage multiplication. The transformer **102** has a turns ratio of N , where N is the ratio of the number of turns in the secondary winding to the number of turns in the primary winding. Transformer **102** has been drawn with a single secondary winding in FIG. 1. However, the forward boost topology disclosed herein could be used with transformers having multiple secondary windings.

The primary side of the transformer **102** has a storage inductance L_{STOR} **114** coupled across the primary winding. The secondary winding of the transformer **102** is coupled to resonant capacitance C_{RES} **116** and resonant inductance $L_{S,RES}$ **118**. This series combination of the secondary winding, $L_{S,RES}$ **118**, and C_{RES} **116** is shunted by a switch **122**, connects to terminal **110** through switch **120**, and connects to terminal **112**. Storage capacitance **124** is coupled across terminals **110, 112**.

Transformer **102** in FIG. 1A has been drawn as an ideal transformer with no parasitic elements. An actual transformer has a number of parasitic elements. Parasitic elements that could be exploited in the operation of a forward boost converter are the transformer’s magnetisation and leakage inductances. FIG. 1B is a schematic diagram illustrating a transformer’s parasitic inductances, including the location of the magnetisation and leakage inductances. The magnetisation inductance L_M is across or in parallel with the primary transformer winding. The current drawn by L_M represents the current required to magnetise the transformer’s core. Transformers are normally designed to maximize the value of L_M since the greater the value of L_M , the smaller the storage current. In a forward boost converter however, a small value of L_M could be useful in storing energy during part of the converter cycle.

L_M may therefore be deliberately decreased in order to increase the energy storage capacity of the transformer. In this respect it is similar to a conventional flyback DC to DC converter. In one embodiment L_M is decreased by inserting a small air gap into the transformer core to increase its reluctance and increase its magnetizing current. In some embodi-

ments of a forward boost converter, the transformer’s magnetisation inductance is sufficiently small that no external storage inductor is required, thereby reducing the number of discrete components. Thus, in some embodiments, storage inductance **114** in FIG. 1A is the intrinsic or parasitic magnetisation inductance of the transformer **102** and is not an external discrete inductor.

Another parasitic transformer element that could be useful in a forward boost converter is the transformer’s parasitic leakage inductance. Leakage inductance results from the imperfect coupling of magnetic flux between the primary and secondary windings. The leakage inductance of a transformer is typically much smaller than its magnetisation inductance. The leakage inductance is represented in FIG. 1B by inductance $L_{S,RES}$ having a value $L_{S,LEAK}$ in series with the secondary winding of the transformer. Alternatively and equivalently the leakage inductance could have been shown as an inductance in series with the primary winding. Transformer leakage inductance is normally considered to be an undesirable parasitic component. In a forward boost converter however, leakage inductance could be advantageously used in series with resonant capacitance C_{RES} **116** to transfer energy. In some embodiments the transformer’s leakage inductance is sufficiently large that no external resonance inductance is required, thereby reducing the number of discrete components. Thus, in some embodiments inductor $L_{S,RES}$ in FIG. 1A is the inherent leakage inductance of the transformer **102** and is not an external discrete inductor.

With reference again to FIG. 1A, switches **104, 120** and **122** may be implemented in a variety of ways, including power MOSFETs (Metal Oxide Semiconductor Field Effect Transistors), BJTs (Bipolar Junction Transistors), IGBTs (Insulated Gate Bipolar Transistors), junction diodes, and/or mechanical relays, for example. In the case of controllable switches, the opening and closing of switches **104, 120** and **122** may be controlled by a control means which is not shown.

The example converter **100** illustrated in FIG. 1A is bi-directional. Depending on the operation of switches **104, 120** and **122**, power may be supplied to terminals **106** and **108** and output at terminals **110** and **112** or alternatively, power can be supplied to terminals **110** and **112** and output at terminals **106** and **108**. The presence of capacitance **124** and capacitance **126** is dependent on the direction of power flow in the example converter **100**. For instance, if the example converter **100** were to only operate with terminals **106** and **108** as the input then capacitance **126** might be unnecessary. Similarly if the example converter **100** were to only operate with terminals **110** and **112** as the input then capacitance **124** might be unnecessary. In practical implementations, however, it is expected that a capacitance **124, 126** would be coupled across each terminal pair **106, 108** and **110, 112**. Capacitance across the input terminal pair, which could be **106, 108** or **110, 112** as noted above, need not necessarily be provided in a converter circuit, but could instead be in the form of a capacitance across the output terminals of a voltage source, for example.

FIG. 2 is a schematic diagram illustrating an example forward boost converter **200** using an anti-phase transformer. In the example converter **200** in FIG. 2, the primary-side terminals (**106** and **108** in FIG. 1A) are the input terminals and the secondary-side terminals (**110** and **112** in FIG. 1A) are the output terminals. Input voltage source **201** of value V_{IN} is applied to the input terminals and load resistance **205** of value R_L is coupled across the output terminals. In this embodiment, the switching elements **120** and **122** (FIG. 1A) are implemented as diodes **D2 220** and **D1 222** respectively.

Using standard notation, the polarity of each winding of the transformer **202** is indicated by the black dots. In the example

embodiment shown in FIG. 2 the windings of the transformer 202 are arranged in anti-phase. However, in another embodiment, the windings might be arranged in phase. Although the transformer 202 is shown in FIG. 2 as a magnetic core transformer, in another embodiment it might be an air core transformer. In the example converter 200, storage inductance L_{STOR} 214 consists entirely of the intrinsic magnetisation inductance of transformer 202 and there is no external storage inductor. Also in this embodiment, resonant inductance $L_{S,RES}$ 218 consists entirely of the intrinsic leakage inductance of transformer 202 of value $L_{S,LEAK}$ and there is no external resonant inductor.

There is optionally a “snubber” circuit 207 coupled across the primary winding of the transformer 202. The snubber circuit 207 is designed to limit the voltage stress on primary switch 204 when it switches off. High voltages can be created across primary switch 204 from the interruption of the current flowing in the storage inductance L_{STOR} 214 when the switch opens. The snubber circuit 207 provides a current path when the switch 204 switches off and limits the voltages across the switch 204. Snubber circuits can be implemented in a variety of ways.

FIG. 3A is a current versus time graph illustrating resonant and storage current waveforms in the primary and secondary windings for the example forward boost converter 200 in FIG. 2. FIG. 3A depicts components of currents in the primary and secondary windings of the transformer 202 for a complete cycle of operation of the example converter 200. FIG. 3A is intended to illustrate general waveform shapes, and units of current and time are arbitrary.

At time $t=0$ under steady load conditions, output capacitance 224 will have been previously charged to a voltage of value V_{OUT} and resonant capacitance C_{RES} 216 will have been previously charged to a voltage close to $V_{IN} \cdot N$ where V_{IN} is the input voltage and the N is the turns ratio of transformer 202. V_{IN} and N are selected, for the purposes of illustration, such that their product is less than V_{OUT} and diode 220 is therefore reverse biased and non-conducting. In one embodiment $V_{IN} \cdot N$ is selected to be about 70% of V_{OUT} . Diode 222 is also reverse biased by the voltage across resonant capacitance C_{RES} 216. The core of the transformer 202 will have been previously completely demagnetised so the currents in the primary and secondary windings of transformer 202 are both zero.

At time $t=0$ primary switch 204 closes to begin its conduction period, and the current in the primary winding (I_P) begins to increase from zero. As shown in FIG. 3A there are two components to the primary current. A first component, I_{P1} , shown in FIG. 3A as “Magnetisation”, is the storage current flowing through storage inductance L_{STOR} 214 and increases linearly with time (t). This current flows through the magnetisation inductance of the transformer 202 and serves to magnetise the core of transformer 202. The storage current creates no corresponding current in the secondary winding or energy transfer to the output side of the circuit. The storage current I_{P1} is given by the formula:

$$I_{P1} = \frac{V_{IN} t}{L_{STOR}}.$$

The magnetic energy E_{MAG} stored in an inductance of value L carrying current i is given by the equation:

$$E_{MAG} = \frac{L_{STOR} i^2}{2}.$$

Hence, the energy (E_{MAG}) stored in the storage inductance L_{STOR} 214 at the end of the conduction period (t_{on}) is:

$$E_{MAG} = \frac{I_{ON}^2 V_{IN}^2}{2 L_{STOR}}.$$

A second component of the primary current (I_{PR}), shown in FIG. 3A as “Primary Resonant”, is due to the resonant LC circuit on the secondary side of the transformer 202, consisting of resonant capacitance C_{RES} 216 coupled in series with resonant inductance $L_{S,RES}$ 218. When primary switch 204 closes, a voltage of V_{IN} appears across the primary winding of transformer 202, creating a voltage of magnitude $V_{IN} \cdot N$, but of opposite polarity, across the secondary winding. The voltage is sufficiently larger than the voltage stored in resonant capacitor C_{RES} 216 to forward bias diode 222. Diode 222 then begins its conduction period and provides a path for current to flow through the secondary winding of transformer 202, capacitor C_{RES} 216 and inductor $L_{S,RES}$ 218. Diode 220 remains reverse biased. FIG. 4 is a schematic diagram illustrating an equivalent circuit for the example forward boost converter 200 in FIG. 2 while the primary switch 204 is closed.

This LC circuit as shown in FIG. 4 has resonant frequency ω_R given by the formula:

$$\omega_R = \frac{1}{\sqrt{L_{S,LEAK} C_{RES}}}.$$

The resonant secondary current (I_{SR}), shown in FIG. 3A as “Secondary Resonant”, charges capacitance C_{RES} 216, causing its voltage to increase by an amount $2 \cdot \Delta V$, equal to the time integral of the secondary current divided by the value of the capacitance. Under steady load the voltage of capacitance C_{RES} 216 is symmetric around $V_{IN} \cdot N$, and thus its voltage at the beginning of the cycle is $V_{IN} \cdot N - \Delta V$.

The resonant secondary current I_{SR} is a time varying sinusoid given by:

$$I_{SR} = \Delta V \sqrt{\frac{C_{RES}}{L_{S,LEAK}}} \sin(\omega_R t).$$

This resonant secondary current I_{SR} is mirrored in the primary winding, producing the second component of primary current I_{PR} . The resonant primary current also has the form of a sinusoid and is given by the formula:

$$I_{PR} = N I_{PEAK} \sin(\omega_R t).$$

Resonant secondary current I_{SR} can only flow while diode 222 is forward biased, hence both I_{SR} and I_{PR} are limited to at most a half cycle of the sinusoid as shown FIG. 3A. In practice the exact shape of the resonant primary current depends on the length of the conduction period of primary switch 204 (t_{on}), set by a switching control means. If the conduction period is longer than the half period of the sinusoid ($T/2$), then a full half cycle of primary and secondary resonant current will appear, as shown in FIG. 3A. If the conduction period is

less than $T/2$ only a portion of the half sinusoid would appear. The period of the sinusoid (T) is given by the formula:

$$T = \frac{2\pi}{\omega_R}$$

Resonant capacitance C_{RES} 216 is sized such that its change in voltage is much less than its steady state voltage of $V_{IN} \cdot N$. In one embodiment $V_{IN} \cdot N$ is 300V and the change in voltage is 6V.

The energy transferred into capacitance C_{RES} 216 during this half sinusoid charging cycle is proportional to $C_{RES} \cdot \Delta V^2$. Thus the amount of energy transferred is independent of the value $L_{S,LEAK}$ of the leakage inductance of transformer 202.

Smaller values of $L_{S,LEAK}$ result in a narrower and higher half sinusoid pulse but the amount of energy transferred is the same. There are some practical limits, however. At very small values of $L_{S,LEAK}$ the current pulse will have such a large amplitude as to cause undesirable conduction losses. Therefore in some embodiments an external inductance may be advantageously added in series with the secondary winding to increase the value of the resonant inductance and reduce the peak of the current pulse.

At very large values of $L_{S,LEAK}$ the width of the half sinusoid will be so long that it will not complete before primary switch 204 is opened to end its conduction period. However, since the pulse width is only proportional to the square root of the leakage inductance, a wide range of values of $L_{S,LEAK}$ can be tolerated.

From the above analysis it will be apparent that the current in the primary winding of the transformer 202 during the conduction phase of primary switch 204 is the superposition of a linear ramp due to the storage inductance and a half sinusoid due to the LC resonant circuit as shown in FIG. 3A.

Energy is stored in the storage inductance L_{STOR} 214 and in the resonant capacitance C_{RES} 216 during the conduction phase of primary switch 204. When switch 204 opens the previously stored energy is transferred into load 205 and output capacitance 224.

When switch 204 opens, the current in the primary winding of the transformer 202 falls to zero and energy stored in storage inductance L_{STOR} 214 and resonant capacitance C_{RES} 216 begins to be transferred into load 205 and output capacitance 224. The effect of storage inductance L_{STOR} 214 can be modeled as an equivalent inductance L'_{STOR} of value $N^2 L_{STOR}$ in the output circuit. When switch 204 opens the voltage across L'_{STOR} rises sufficiently to forward bias diode 220 and begin its conduction period, to thereby allow current to flow into load 205 and capacitance 224. Diode 222 becomes reverse biased to end its conduction period. Ideally, no primary current flows since switch 204 is open.

FIG. 5 is a schematic diagram illustrating an equivalent circuit for the example forward boost converter 200 in FIG. 2 when the primary switch 204 opens. As noted above, L'_{STOR} 515 represents an equivalent inductance that models the effect of storage inductance L_{STOR} 214.

Capacitance 224 is sized relative to inductance L'_{STOR} 515 such that the output voltage of the converter V_{OUT} will not change significantly. Therefore, the voltage on that side of inductance L'_{STOR} 515 is effectively clamped to a diode drop above V_{OUT} . Similarly the other side of inductance L'_{STOR} 515 is clamped to about $V_{IN} \cdot N$ by capacitance C_{RES} 216, resulting in a voltage of about $V_{OUT} - V_{IN} \cdot N$ across inductance L'_{STOR} 515. This voltage is mirrored across to the primary winding of transformer 202, creating a voltage of $V_{OUT}/$

$N - V_{IN}$ across it. The voltage across primary switch 204 (V_D) therefore rises to about V_{OUT}/N .

The secondary winding current waveform is shown in FIG. 3A. The secondary winding current decreases approximately linearly to zero as the energy stored in inductance L'_{STOR} 515 is transferring into capacitance 224 and load 205. The core of transformer 202 is also demagnetizing during this operation. The voltage of resonant capacitance C_{RES} 216 also decreases to $V_{IN} \cdot N - \Delta V$ as its stored energy is transferred into load 205 and output capacitance 224 by the current in inductance L'_{STOR} 515, creating an additional "boost" of energy.

Once the core of transformer 202 is completely demagnetized, secondary current flow ceases and diode 220 becomes reverse biased, to end its conduction period, while diode 222 remains reverse biased. It should be noted that the equivalent circuit in FIG. 5 contains a series LC resonance consisting of inductance L'_{STOR} 515, inductance $L_{S,RES}$ 218 and capacitance C_{RES} 216. However, since the value of the inductance L'_{STOR} 515 is so much larger than the value of the inductance $L_{S,RES}$ 218 the resonant behavior cannot be seen on the time scale shown in FIG. 3A, and the current waveform appears to be linear.

After the core of transformer 202 is demagnetized, the primary and secondary windings of the transformer are both open. The primary winding is opened by the opening of primary switch 204 and the secondary winding is opened by diodes 220 and 222 that are both reverse biased. Parasitic capacitances will typically exist across the diodes 220, 222 and the switch 204, however. If switch 204 is implemented with a power MOSFET for example, then its parasitic capacitance is substantially the MOSFET source drain capacitance. These parasitic capacitances together with the parasitic capacitance of the windings of transformer 202 create a parallel LC resonant circuit and the transformer magnetisation inductance will resonate. FIG. 6 is a schematic diagram illustrating an equivalent circuit for the example forward boost converter 200 in FIG. 2 during the transformer resonance period.

There is no energy transfer between input and output during this resonance period. When the input power to the example converter 200 is very low, it may be advantageous to keep the converter in this resonant state rather than have it continue to switch. This minimizes switching losses and maintains power efficiency. To exit the resonant state and begin the next cycle, the primary switch 204 could be closed when the voltage across it is at an oscillatory minimum. This further minimizes the switching loss.

FIG. 3B is a current versus time graph illustrating simulated primary and secondary current waveforms for the example forward boost converter 200 in FIG. 2, produced using the commercial circuit simulator SIMPLIS. Current waveforms observed during actual operation of a forward boost converter as disclosed herein may exhibit different characteristics, which are implementation-dependent.

A unique aspect of the example converter 200 is that energy is transferred from the converter's input to its output on both the conducting and non-conducting periods of the primary switch 204. During the time primary switch 204 is closed, energy is transferred into resonant capacitance C_{RES} 216 and stored in the storage inductance L_{STOR} 214. During the time primary switch 204 is open, energy is transferred from resonant capacitance C_{RES} 216 and storage inductance L_{STOR} 214 to the output. A potential advantage of a forward boost converter as disclosed herein is thus increased efficiency and reduced losses since energy is transferred on both switching cycles. The stress on transformer 202 is less than in a traditional flyback converter since less energy needs to be stored in

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the transformer. Transformer 202 can be made smaller, reducing cost and size of the example converter 202 relative to flyback or forward converter designs. Component count can also be reduced since intrinsic elements of the transformer 202 like magnetisation and leakage inductance are used instead of dedicated external components.

Switching losses from opening primary switch 204 are also reduced compared to a conventional flyback converter. Referring to FIG. 3A, the current flowing through primary switch 204 when it opens consists only of the magnetisation or storage current. The resonant current is zero and the primary current is less than its maximum value. This compares favorably to a conventional flyback converter in which the primary current is a maximum when the primary switch opens. In one embodiment the opening of primary switch 204 is timed to coincide with the termination of the half sinusoid pulse, minimizing the switching losses.

The reverse bias stress on diodes 220 and 222 is also reduced in this topology. The reverse voltage on the diodes 220 and 222 is never more than V_{OUT} , whereas in a conventional flyback converter the reverse bias on the diodes is $V_{OUT} + V_{IN} * N/D$, where D is the duty cycle of the primary switch.

Forward boost converter design is not restricted to the use of transformers with anti-phase windings. FIG. 7 is a schematic diagram illustrating an example forward boost converter circuit 800 using a transformer with in-phase windings. Input voltage source 801 of value V_{in} is applied to the input terminals and load resistance 805 is connected to the output terminals. In this example transformer 802 is a magnetic core transformer and storage inductance L_{STOR} 814 consists entirely of the magnetisation inductance of transformer 802 and there is no external storage inductor. Also in this example, resonant inductance $L_{S,RES}$ 818 consists entirely of the intrinsic leakage inductance of transformer 802 of value $L_{S,LEAK}$ and there is no external resonant inductor. Like the example converter 200 in FIG. 2, the example converter 800 also includes a primary switch 804, a resonant capacitance $C_{S,RES}$ 846, diodes 820, 822, and an output capacitance 824, operatively coupled together as shown. The operation of the example converter 800 is conceptually similar to the example converter 200 in FIG. 2, which includes an anti-phase transformer. There is optionally a "snubber" circuit 807 coupled across the primary winding of the transformer 802 to limit the voltage stress on primary switch 804 when it switches off.

FIG. 10 is a graph illustrating a simulated voltage waveform across the primary switch 804 and simulated primary and secondary current waveforms for the example forward boost converter 800 in FIG. 7. The waveforms were obtained using the commercial circuit simulator SIMPLIS. Waveform characteristics may vary with different simulation parameters or actual converter implementations.

In steady state load conditions, just prior to the closing of primary switch 804, output capacitance 824 is charged to a voltage of V_{OUT} and resonant capacitance C_{RES} 816 is charged to slightly below $V_{OUT} - V_{IN} * N$. V_{IN} and N are selected such that their product is less than V_{OUT} and diode 820 is therefore reverse biased and non-conducting. Diode 822 is also reverse biased and non-conducting and no current flows in the primary and secondary windings of the transformer 802. When primary switch 804 closes to begin its conduction period, current begins to flow in the primary winding. As in the previous embodiment, the primary current has two components, the storage current flowing in storage inductance L_{STOR} 814 and a resonant current. The storage current serves to magnetise the core of transformer 802, and therefore there is no corresponding current in the secondary

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winding and there is no energy transfer to the output side of the example converter 800. The storage current is a approximately a linear time-dependent ramp.

When primary switch 804 closes, a voltage appears across the secondary winding of the transformer 802, forward biasing diode 820 to begin its conduction period. This contrasts with the previous embodiment 200 of FIG. 2 in which diode 222 became forward biased on the closing of the primary switch. A resonant current flows in the LC resonant circuit composed of the series combination of the leakage inductance $L_{S,RES}$ 818, capacitance C_{RES} 816, and output capacitance 824.

FIG. 8 is a schematic diagram illustrating an equivalent circuit for the example forward boost converter 800 in FIG. 7 while the primary switch 804 is closed.

Unlike the anti-phase case (FIG. 2), output capacitance 824 is a part of the resonant circuit. Output capacitance 824, however, is chosen to be significantly larger than resonant capacitance C_{RES} 816 such that the series combination of the two capacitances results in a total value of capacitance not much smaller than the value of C_{RES} 816. In one embodiment capacitance C_{RES} 816 has a value of 220 nF and output capacitance 824 has a value of 1 uF. As in the anti-phase case (FIG. 2), only a half sinusoid pulse of resonant current flows. Energy is therefore transferred into load 805 and output capacitance 824 from capacitance C_{RES} 816 during the conduction period of the primary switch 804 in this configuration. This contrasts with the anti-phase configuration of the previous embodiment (FIG. 2) where energy was transferred into the load from capacitance C_{RES} 816 during the non-conducting period of the primary switch 804.

The storage current in the primary winding continues to flow until the primary switch 804 is opened to end its conduction period. The primary current is a superposition of the storage and resonant currents as discussed above.

At the end of its conduction period, primary switch 804 opens and the current in the primary winding falls to zero. The energy stored in the storage inductance L_{STOR} 814 begins to be transferred into the secondary side. FIG. 9 is a schematic diagram illustrating an equivalent circuit for example forward boost converter 800 in FIG. 7 when the primary switch 804 opens. The effect of the storage inductance L_{STOR} 814 can be modeled as equivalent inductance L'_{STOR} 1015 of value $N^2 * L_{STOR}$ in the output circuit.

When primary switch 804 opens, the voltage induced across the equivalent inductance L'_{STOR} 1015 is in the opposite direction to the voltage across C_{RES} 816. It forward biases diode 822 to begin its conduction period, while diode 820 is reverse biased to end its conduction period.

The secondary winding current waveform is shown in FIG. 10. The secondary winding current magnitude decreases almost linearly to zero as the energy stored in L'_{STOR} 1015 is transferring into resonant capacitor C_{RES} 816. The core of the transformer 802 is also demagnetizing during this operation. The voltage of capacitor C_{RES} 816 increases to slightly above $V_{OUT} - V_{IN} * N$ as it stores energy from the inductor current. It should be noted that the equivalent circuit in FIG. 9 is a series LC resonant circuit. However, the value of the equivalent inductance L'_{STOR} 1015 is so large that the sinusoidal nature of the current cannot be clearly seen on the time scale in FIG. 10 and the current magnitude appears to decrease substantially linearly.

When the core of the transformer 802 is completely demagnetised, secondary current flow ceases and diode 822 becomes reverse biased to end its conduction period.

As in the previous embodiment the primary and secondary windings of transformer 802 are now both open and the

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storage inductance **814** will resonate with the parasitic capacitance elements of the circuit. The resonance creates a sinusoidally varying voltage across primary switch **804** as illustrated in FIG. 10.

As in the previous embodiment, there is no energy transfer between the input and the output during this resonant period. When the input power to the example converter **800** is very low, the converter could be kept in this resonant period state to maintain power efficiency. This can be achieved by reducing the switching frequency of the example converter **800** while maintaining the same ON period for the primary switch **804**.

This forward boost topology using an in-phase transformer is conceptually similar to the topology with anti-phase transformer. Energy is transferred from the input stage of each example converter **200**, **800** to its output stage on both the conducting and non-conducting periods of each primary switch **204**, **804**. During the conducting period of primary switch **804** energy is stored in the storage inductance **814** but energy is also transferred from resonant capacitance C_{RES} **816** into load **805** and output capacitance **824**. During the non-conducting period of primary switch **804** energy is transferred from the storage inductance **814** to capacitance C_{RES} **816**. This embodiment, like the previous embodiment, may provide for increased efficiency, smaller transformer sizing, reduced component count and reduced stress on the diodes.

The ability to operate a power converter in reverse could be useful, for example, in applications involving electrical storage such as batteries or fuel cells in which the storage will supply power for certain periods and accept power for other periods. FIG. 11 is a schematic diagram illustrating an example forward boost converter **1200** operated in the reverse direction to the example forward boost converter **200** in FIG. 2. An anti-phase transformer **1202** is used with a primary to secondary turns ratio of $1/N$ and a magnetic core in the example shown, although an in-phase transformer, or an anti-phase or in-phase air core transformer might instead be used. Storage inductance L_{STOR} **1214** consists entirely of the magnetisation inductance of transformer **1202** and there is no external storage inductor. Also in this example, resonant inductance $L_{P,RES}$ **1218** consists entirely of the intrinsic leakage inductance of transformer **1202** with value $L_{S,LEAK}$ and there is no external resonant inductor.

In accordance with convention, the winding of transformer **1202** at the input side of the example forward boost converter **1200** is referred to as the primary winding and the winding at the output side is referred to as the secondary winding. However, the example forward boost converter **1200** illustrates reverse operation relative to the example forward boost converter **200** in FIG. 2, and therefore L_{STOR} , which is the magnetisation inductance of transformer **1202**, is shown at the output side winding in FIG. 11, and $L_{P,RES}$, which is the leakage inductance of transformer **1202**, is shown at the input side winding.

The example forward boost converter **1200** also includes an input source **1201**, capacitance **1224**, C_{RES} **1216**, capacitance **1226**, and resistive load **1205**, coupled together as shown. Controllable switches **1220**, **1222** replace diodes **220/820**, **222/822** in previous embodiments, and diode **1204** replaces primary switch **204**, **804** in previous embodiments.

Input voltage source **1201** of value V_{in} is applied to the input terminals and load resistance **1205** is connected to the output terminals.

Resonant capacitance C_{RES} **1216** is now located on the input side of the circuit rather than the output side, and resonates with resonant inductance $L_{P,RES}$ **1218**. Diode **1204** performs a function similar to primary switch **204** in the example forward boost converter **200**, but is located on the output side

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of the example converter **1200**. Although diode **1204** is used in this embodiment, another type of switching device such as a MOSFET or a relay could instead be used. High side switch **1220** and low side switch **1222** perform functions similar to diodes **220** and **222** of the example converter **200**. Switches **1220** and **1222** can be implemented in a variety of ways. In one embodiment, they are N type MOSFET and bypass diodes, and **1221** and **1223** are the parasitic diodes formed by the P-type body and N type drain of each MOSFET.

FIG. 13 is a graph illustrating simulated high side and low side switch drive voltages "Drive HS" and "Drive LS", respectively, and current waveforms for the high side switch **1220**, low side switch **1222** and secondary winding for the example forward boost converter **1200** in FIG. 11.

At the start of the operating cycle under steady state load conditions switches **1220** and **1222** are open, capacitance **1226** is charged to voltage V_{OUT} , capacitance C_{RES} **1216** is charged to about $V_{IN}-V_{OUT}*N$, diode **1204** is reverse biased and the core of transformer **1202** is demagnetised.

When switch **1220** is closed to begin its conduction period, input voltage source **1201** is connected across the primary winding of transformer **1202**. A current begins to flow through high side switch **1220** into the primary winding and a reflected voltage appears across the secondary winding. No secondary current flows since transformer **1202** is anti-phase and keeps diode **1204** reverse biased.

FIG. 12A is a schematic diagram illustrating an equivalent circuit for the example forward boost converter **1200** in FIG. 11 while high side switch **1220** closes. Reflected storage inductance L'_{STOR} **1315** is in series with resonant inductance $L_{P,RES}$ **1218** and resonance capacitance C_{RES} **1216**. The value of the reflected storage inductance L'_{STOR} **1315** is N^2*L_{STOR} . The primary current increases nearly linearly with time since the resonant frequency of the series combination of $L_{P,RES}$ **1218**, L'_{STOR} **1315** and C_{RES} **1216** is low relative to the time scale of the switching. In one embodiment L_{STOR} **1214** has a value of 7.2 uH, C_{RES} **1216** has a value of 220 nF, $L_{S,LEAK}$ is 90 nH and N is 9. This primary current charges resonant capacitance C_{RES} **1216** and stores energy in the storage inductance L_{STOR} **1214**. No current flows in the secondary winding of transformer **1202** since diode **1204** is reverse biased.

When high side switch **1220** is opened to end its conduction period, the current through it stops and the energy stored in storage inductance **1214** begins to transfer to the output. FIG. 12B is a schematic diagram illustrating an equivalent circuit for the input portion of the example forward boost converter **1200** in FIG. 11 when high side switch **1220** opens. The primary winding current does not fall immediately to zero since the diode **1223** across switch **1222** briefly provides a path for continued flow. This current appears as the small current spike in the low side switch current waveform in FIG. 13 co-incident with the falling edge of the DRIVE_HS voltage.

In the output portion of the example converter **1200**, the voltage across storage inductance L_{STOR} **1214** forward biases diode **1204** to begin its conduction period. FIG. 12C is a schematic diagram illustrating an equivalent circuit for the output portion of the example forward boost converter **1200** in FIG. 11 when high side switch **1220** opens. Secondary current flows as storage inductance L_{STOR} **1214** discharges into load **1205** and output capacitance **1226**. The current flow decreases approximately linearly, as shown in FIG. 13.

After storage inductance L_{STOR} **1214** has discharged, low side switch **1222** is closed to begin its conduction period, forming a series LC resonant circuit from resonant capacitance C_{RES} **1216** and resonant inductance $L_{P,RES}$ **1218** in the primary winding. FIG. 12D is a schematic diagram illustrat-

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ing an equivalent circuit for the example forward boost converter **1200** in FIG. **11** when low side switch **1222** is closed. Resonant inductance $L_{P,RES}$ **1218** and resonant capacitance C_{RES} **1216** form a series resonant circuit with resonant frequency ω_R of value

$$\omega_R = \frac{1}{\sqrt{L_{S,LEAK} C_{RES}}}.$$

Resonant currents flow in both the primary and secondary windings. As in the previous examples, a maximum of a half sinusoid current pulse will flow before diode **1204** becomes reverse biased to end its conduction period.

Capacitance C_{RES} **1216** also supplies current into storage inductance L_{STOR} **1214**, energizing it. This storage current is approximately a linear ramp due to the much greater inductance of L_{STOR} **1214**. The current through low side switch **1222** is therefore the sum of the resonant and storage currents and is a half sinusoid pulse impressed on a linear ramp. The storage current is not reflected into the secondary and the secondary current therefore consists only of the resonant current and has the shape of a half sinusoid. The secondary current ceases once the half sinusoid completes and diode **1204** is reverse biased.

The storage current through low side switch **1222** continues until it is opened to end its conduction period. The voltage across storage inductance L_{STOR} **1214** then reverses to maintain the current and forward biases diode **1221** across high side switch **1222**. FIG. **12E** is a schematic diagram illustrating an equivalent circuit for the input portion of the example forward boost converter **1200** in FIG. **11** when low side switch **1222** is opened.

The storage current decreases linearly to zero and diode **1221** becomes reverse biased. No secondary current flows since diode **1204** is reverse biased.

Both windings of the transformer **1202** are now open and the transformer resonates. As in the previous embodiment no energy transfer takes place between the input and output of the example converter **1200** during this resonant period. When the input power to the example converter **1200** is very low, the converter could be kept in this resonant period state to maintain power efficiency. This can be achieved by reducing the switching frequency of the example converter **1200** while maintaining the same ON periods for low side switch **1222** and high side switch **1220**.

The foregoing describes particular embodiments, but is not meant to be limiting. For example, although an anti-phase transformer was used in the example of reverse operation of a forward boost converter, an in-phase transformer could be used. Although the parasitic leakage inductance of the transformer **1202** is used to form a resonant circuit with resonant capacitance **1216**, a dedicated and separate, discrete inductor could be used for that purpose. Although the magnetisation inductance of the transformer **1202** is used as the storage inductance **1214**, a dedicated and separate, discrete inductor could be used for this purpose.

Embodiments of the forward boost converter are not limited to converters employing transformers between their inputs and outputs. FIG. **14** is a schematic diagram illustrating another example forward boost converter **1500**. There is no transformer and the inductive elements are provided by inductors **1514**, **1518**. The operation of the example converter **1500** is, however, similar to the previously described forward boost converters. Like the other embodiments, described above, the example forward boost converter **1500**, in addition

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to the inductive elements **1514**, **1518**, includes an input source **1501**, capacitive elements **1526**, **1516**, **1524**, and switching elements **1504**, **1520**, **1522**. A resistive load is also shown at **1505**, and an optional snubber is shown at **1507**.

At the beginning of an operating cycle, capacitance **1524** is charged to V_{OUT} , capacitance C_{RES} **1516** is charged to approximately $V_{IN}-V_{OUT}$ and diodes **1522** and **1520** are reverse biased. When switch **1504** closes to begin its conduction period, a storage current flows into storage inductance L_{STOR} **1514** from supply **1501**. Diode **1522** also becomes forward biased to begin its conduction period, and a resonant current flows through the series combination of resonant capacitance C_{RES} **1516** and resonant inductance L_{RES} **1518**. As in the previous examples the resonant current consists of a half sinusoidal current pulse while the storage current consists of a linear ramp. No energy is transferred to load **1505**. When switch **1504** opens to end its conduction period, the voltage across storage inductance L_{STOR} **1514** reverses, diode **1520** becomes forward biased to begin its conduction period, and diode **1522** becomes reverse biased to end its conduction period. The current in L_{STOR} **1514** decreases approximately linearly to zero as the energy stored in it is transferring into output capacitance **1524** and load **1505**. The energy stored in the resonant capacitance C_{RES} **1516** is also transferred into load **1505** and output capacitance C_{RES} **1524** by the storage inductor current, creating the characteristic “boost” in energy.

FIG. **15** is a schematic diagram illustrating an alternate example of a forward boost converter **1600** which functions in substantially the same manner as **1500**. The example forward boost converter **1600** includes an input voltage source **1601**, capacitive elements **1626**, **1616**, **1624**, inductive elements **1614**, **1618**, switching elements **1604**, **1622**, **1620**, resistive load **1605**, and optionally a snubber **1607**, coupled together as shown.

At the beginning of a cycle switch **1604** closes to begin its conduction period and storage current flows into storage inductance L_{STOR} **1614**. Diode **1620** becomes forward biased to begin its conduction period, and resonant current flows through inductance **1618** and capacitance **1616** into capacitance **1624** and load **1605**. After the half sinusoid pulse completes, switch **1604** opens to end its conduction period, diode **1622** becomes forward biased to begin its conduction period, and diode **1620** becomes reverse biased to end its conduction period, placing storage inductance **1614** in series with resonant inductance **1618** and resonant capacitance **1616**. The current in L_{STOR} **1614** decreases approximately linearly to zero as the resonant capacitance **1616** charges up to V_{OUT} .

FIG. **16** is a schematic diagram illustrating another example of a forward boost converter **1700** which functions in substantially the same manner as the example forward boost converter **1500**. The example forward boost converter **1700** includes an input voltage source **1701**, capacitive elements **1726**, **1716**, **1724**, inductive elements **1714**, **1718**, switching elements **1704**, **1722**, **1720**, resistive load **1705**, and optionally snubber **1707**, coupled together as shown. In this example converter **1700**, storage inductance **1714** charges through resonant inductance **1718** during the conduction period of switch **1704** and discharges through resonant capacitance **1716** during the non-conducting period of switch **1714**.

Forward boost converters as disclosed herein are also not limited to transformers with a single secondary winding. FIG. **17** is a schematic diagram illustrating an example forward boost converter **1800** in reverse operation with multiple secondary windings. The primary side of the example converter **1800** includes an input source **1801**, capacitive elements **1826**, **1816**, inductive elements **1818** and **1814**, high side and

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low side switches **1820**, **1822** with parasitic diodes **1821**, **1823**, and the primary winding of the transformer **1802**. Each secondary winding is coupled to a diode **1830**, **1832**, **1834**, **1836** and an output capacitive element **1831**, **1833**, **1835**, **1837**. The multiple secondary windings produce output voltages V_{OUT1} , V_{OUT2} , V_{OUT3} and V_{OUT4} in the example shown. Output voltage V_{OUT3} is across the entire split secondary winding shown in FIG. 17, and V_{OUT4} is across the lower portion of the split secondary winding. The output terminal pairs for output voltages V_{OUT3} and V_{OUT4} thus share one common terminal, specifically the lower terminal of the split secondary winding in this example.

In FIG. 17, the storage inductance L_{STOR} **1814** is shown across the primary winding of transformer **1802** according to common convention and for simplicity. It will be understood, however, that L_{STOR} will appear as a reflected inductance in each of the secondary windings and is reflected by the square of the ratio of their respective number of turns.

FIGS. 18 to 22 are schematic diagrams illustrating equivalent circuits for example forward boost converter operating states.

FIGS. 18 to 22 are schematic diagrams illustrating equivalent circuits for example states of the forward boost converter. The equivalent circuit shown in FIG. 18 for a state 1 is common to the example embodiments shown in FIGS. 2, 7, 11, and 14 to 17, in which a first inductance is coupled to a first switch in a first circuit path across a first terminal pair. The storage inductance L_{STOR} is coupled across one of the converter's two terminal pairs by a conducting switch, labelled in FIG. 18 as "1st Switch", during a conduction period of that switch. In this state, and in forward operation of a forward boost converter, energy is transferred into the storage inductance L_{STOR} from the terminals. In this state, but in reverse operation, energy is transferred out of the storage inductance L_{STOR} into the terminals.

Embodiments of the forward boost converter as disclosed herein also use a combination of states 2 and 4 (FIGS. 19 and 21) or states 3 and 5 (FIGS. 20 and 22).

Considering forward operation as an example, in state 2, a circuit path including the resonant capacitance C_{RES} and storage inductance L_{STOR} is completed by a conducting low side switch, labelled as "3rd Switch" in FIG. 19, during its conduction period and energy flows into the resonant capacitance from the storage inductance. In state 4, a circuit path including the resonant capacitance C_{RES} and resonant inductance $L_{P,RES}$ is coupled to the second terminal pair during a conduction period of a high side switch, labelled as "2nd Switch" in FIG. 21, and energy is transferred from the resonant capacitance into the terminal pair.

In state 3, a circuit path including the storage inductance L_{STOR} and the resonant capacitance C_{RES} is coupled across the second terminal pair by the conducting high side switch during its conduction period. In this state, in forward operation, energy is transferred into the terminals from the storage inductance L_{STOR} and the resonant capacitance C_{RES} . In this state, in reverse operation, energy is transferred from the terminals into the storage inductance L_{STOR} and resonant capacitance C_{RES} . Although the resonant inductance $L_{P,RES}$ is shown in series with the resonant capacitance C_{RES} and storage inductance L_{STOR} in FIG. 20, this is optional and does not occur in all embodiments, such as in the example converter shown in FIG. 16.

In state 5, a circuit path including the resonant inductance $L_{P,RES}$ and resonant capacitance C_{RES} is completed by the conducting low side switch during its conduction period and the inductance and capacitance resonate for a maximum of a half cycle.

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What has been described is merely illustrative of the application of principles of embodiments. Other arrangements and methods can be implemented by those skilled in the art without departing from the scope of the present invention.

For example, although the load in the preceding embodiments is shown as resistive, forward boost converters as disclosed herein are not restricted to driving resistive loads.

While the invention has been illustrated and described in detail in the drawings and foregoing description, the same is to be considered as illustrative and not restrictive in character, it being understood that only the preferred embodiment has been shown and described and that all changes, equivalents, and modifications that come within the spirit of the inventions defined by following claims are desired to be protected. All publications, patents, and patent applications cited in this specification are herein incorporated by reference as if each individual publication, patent, or patent application were specifically and individually indicated to be incorporated by reference and set forth in its entirety herein.

We claim:

1. A switching mode power converter coupled between a first terminal pair and a second terminal pair, the converter comprising:

a first inductance coupled to a first switch in a first circuit path across said first terminal pair, said first switch coupling said first inductance across said first terminal pair during a conduction period of said first switch;

a capacitance coupled to a second inductance in a second circuit path, and to said first inductance in a third circuit path;

a second switch and a third switch, said second switch operated to be in a non-conducting state when said third switch is in a conducting state and said third switch operated to be in a non-conducting state when said second switch is in a conducting state,

said second switch completing a circuit between said second terminal pair and one of:

said second circuit path and said third circuit path during a conduction period of said second switch,

said third switch completing the other of: said second circuit path and said third circuit path during a conduction period of said third switch.

2. The switching mode power converter of claim 1, wherein said first terminal pair and said second terminal pair share a common terminal.

3. The switching mode power converter of claim 1, said first terminal pair comprising an input terminal pair, said second terminal pair comprising an output terminal pair, said capacitance transferring energy to said output terminal pair during said conduction period of said second switch.

4. The switching mode power converter of claim 3, said first inductance being coupled to said capacitance and transferring energy to said output terminal pair during said conduction period of said second switch.

5. The switching mode power converter of claim 3, said first inductance being coupled to said capacitance and transferring energy to said capacitance during said conduction period of said third switch.

6. The switching mode power converter of claim 1, said first terminal pair comprising an output terminal pair, said second terminal pair comprising an input terminal pair, said first inductance transferring energy to said output terminal pair during said conduction period of said first switch.

7. The switching mode power converter of claim 6, said first inductance being coupled to said capacitance and storing energy from said input terminal pair during said conduction period of said second switch.

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8. The switching mode power converter of claim 7, said first inductance and said capacitance transferring energy to said output terminal pair during said conduction period of said first switch.

9. The switching mode power converter of claim 1, said first inductance being coupled to said capacitance through a transformer.

10. The switching mode power converter of claim 9, said first inductance comprising magnetisation inductance of said transformer.

11. The switching mode power converter of claim 9, said second inductance comprising leakage inductance of said transformer.

12. The switching mode power converter of claim 9, said transformer comprising a first winding coupled across said first inductance and a second winding coupled in said second circuit path with said capacitance and said second inductance.

13. The switching mode power converter of claim 12, said transformer further comprising a third winding.

14. The switching mode power converter of claim 10, said transformer further comprising a third winding, said switching mode power converter further comprising a fourth switch coupled to said third winding and a reflected inductance of said transformer magnetization inductance in a fourth circuit path across a third terminal pair, said fourth switch having said conduction period of said third switch and coupling said fourth circuit path across said third terminal pair during said conduction period of said third and fourth switches.

15. The switching mode power converter of claim 10, said transformer further comprising a third winding, said switching mode power converter further comprising: a fourth switch coupled to said third winding and a first reflected inductance of said transformer magnetisation inductance in a fourth circuit path across a third terminal pair, said fourth switch having said conduction period of said third switch and coupling said fourth circuit path across said third terminal pair during said conduction period of said third and fourth switches;

a fifth switch coupled to a portion of said third winding and a second reflected inductance of said transformer magnetization inductance in a fifth circuit path across a fourth terminal pair, said fifth switch having said conduction period of said third switch and coupling said fifth circuit path across said fourth terminal pair during said conduction period of said third and fifth switches.

16. The switching mode power converter of claim 1, further comprising one or more of:

a capacitance coupled across said first terminal pair; and a capacitance coupled across said second terminal pair.

17. The switching mode power converter of claim 9 wherein said first inductance resonates during a non-conduction period of said first switch, said second switch and said third switch to produce a time varying voltage across said first switch.

18. The switching mode power converter of claim 17 wherein said first switch enters said conduction period while said voltage is at an oscillatory minimum.

19. A method of operation of a switching mode power converter coupled between a first terminal pair and a second terminal pair, the converter comprising: a first inductance coupled to a first switch in a first circuit path across said first terminal pair; a capacitance coupled to a second inductance in a second circuit path, and to said first inductance in a third current circuit path; a second switch and a third switch, said

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second switch operated to be in a non-conducting state when said third switch is in a conducting state and said third switch operated to be in a non-conducting state when said second switch is in a conducting state, said method comprising:

closing of said first switch to couple said first inductance across said first terminal pair during a conduction period of said first switch;

closing of said second switch to complete a circuit between said second circuit path or said third circuit path and said second terminal pair during a conduction period of said second switch so as to transfer energy into said second terminal pair; and

closing of said third switch to complete said second circuit path or said third circuit path during a conduction period of said third switch so as to transfer energy into said capacitance.

20. Apparatus comprising:

a switching mode power converter coupled between a first terminal pair and a second terminal pair, the converter comprising: a first inductance coupled to a first switch in a first circuit path across said first terminal pair; a capacitance coupled to a second inductance in a second circuit path, and to said first inductance in a third current circuit path; a second switch and a third switch, said second switch operated to be in a non-conducting state when said third switch is in a conducting state and said third switch operated to be in a non-conducting state when said second switch is in a conducting state;

means for closing said first switch to couple said first inductance across said first terminal pair during a conduction period of said first switch;

means for closing said second switch to complete a circuit between said second circuit path or said third circuit path and said second terminal pair during a conduction period of said second switch so as to transfer energy into said second terminal pair; and

means for closing said third switch to complete said second circuit path or said third circuit path during a conduction period of said third switch so as to transfer energy into said capacitance.

21. The method of claim 19, wherein:

the closing of said first switch to couple said first inductance across said first terminal pair during a conduction period of said first switch transfers energy by a substantially linearly varying current;

the closing of said second switch to complete a circuit between said second circuit path and said second terminal pair during a conduction period of said second switch transfers energy by a substantially half sinusoidal current pulse;

the closing of said second switch to complete a circuit between said third circuit path and said second terminal pair during a conduction period of said second switch transfers energy by a substantially linearly varying current;

the closing of said third switch to complete said second circuit path during a conduction period of said third switch transfers energy into said capacitance by a resonance of said capacitance for a half sinusoid cycle; and the closing of said third switch to complete said third circuit path during a conduction period of said third switch transfers energy into said capacitance by a substantially linearly varying current.

22. The method of claim 21, comprising:

operation of said switching mode power converter in a forward direction wherein energy is transferred from said first terminal pair to said second terminal pair.

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23. The method of claim **21**, comprising:
operation of said switching mode power converter in a
reverse direction wherein energy is transferred from said
second terminal pair to said first terminal pair.

* * * * *

UNITED STATES PATENT AND TRADEMARK OFFICE
CERTIFICATE OF CORRECTION

PATENT NO. : 9,219,421 B2
APPLICATION NO. : 13/657171
DATED : December 22, 2015
INVENTOR(S) : Gabriel Scarlatescu

Page 1 of 1

It is certified that error appears in the above-identified patent and that said Letters Patent is hereby corrected as shown below:

In the Claims:


In claim 15, Column 19, line 43, please change "magnetization" to --magnetisation--.

In claim 15, Column 19, line 43, please change "in a in a" to --in a--.

In claim 16, Column 19, line 51, please change "a capacitance" to --a second capacitance--.

In claim 16, Column 19, line 52, please change "a capacitance" to --a third capacitance--.

Signed and Sealed this
Seventeenth Day of May, 2016

A handwritten signature in black ink, reading "Michelle K. Lee". The signature is fluid and cursive, with the first letters of each name being capitalized and prominent.

Michelle K. Lee
Director of the United States Patent and Trademark Office